UNCLASSIFIED

AD NUMBER
AD820675
LIMITATION CHANGES
TO: **Approved for public release; distribution is unlimited.**
FROM: **Distribution authorized to U.S. Gov't. agencies and their contractors; Critical Technology; SEP 1967. Other requests shall be referred to Rome Air Development Center, Attn: EMlI, Griffiss AFB, NY I3441-5700. This document contains export-controlled technical data.**
AUTHORITY
RADC ltr, 17 sep 1971

THIS PAGE IS UNCLASSIFIED

AD820675

RADC-TR-67-256
Interim Report

SIGNAL PROCESSING TEST FACILITY (PRELIMINARY DESIGN STUDY)

Prepared By
Techniques Branch
Surveillance and Control Division

TECHNICAL REPORT NO. RADC-TR- 67-256
September 1967

This document is subject to special export controls and each transmittal to foreign governments, foreign nationals or representatives thereto may be made only with prior approval of RADC (EMLI), GAFB, N.Y. 13440

Rome Air Development Center
Air Force Systems Command
Griffiss Air Force Base, New York

Best
Available
Copy

SIGNAL PROCESSING TEST FACILITY (PRELIMINARY DESIGN STUDY)

Prepared By
Techniques Branch
Surveillance and Control Division

See Form 1473

AFLC, GAFB, N.Y., 9 Oct 67-140

FOREWORD

This report represents the results of an in-house design effort by personnel of the Techniques Branch of the Surveillance and Control Division. This effort was devoted to the design of a very high resolution sensor using the best available techniques and a study of the feasibility of converting available equipment into a test vehicle utilizing these techniques.

The following RADC individuals have participated in establishing the design of the radar modifications and have made contributions to this report.

Mr. Thomas Maggio	EMATS
Mr. Clarence F. Silfer	EMATS
Mr. Douglas Mather	EMATA
Mr. Franklin Bradley	EMATT
Mr. John Potenza	EMATA
Mr. Vincent Vannicola	EMATE
Mr. Ronald Vandivier	EMATE
Mr. Patsy Romanelli	EMATE
Mr. Dirk Bussey	EMATE
Mr. Daniel Budzynski	EMATS
Mr. Howard Beard	EMATP
Mr. Bobby Gray	EMATP
Mr. John E. Schneider	EMEAM
Mr. George Bennison	EMCTI

Also acknowledged are the guidance and encouragement of Mr. Arthur J. Frohlich and Dr. Fred I. Diamond.

This report has been reviewed and is approved.

Approved: ARTHUR J. FROHLICH
Chief, Techniques Branch
Surveillance and Control Division

Approved: THOMAS S. BOND, JR.
Colonel, USAF
Chief, Surveillance and Control Division

FOR THE COMMANDER: IRVING J. GABELMAN
Chief, Advanced Studies Group

ABSTRACT

This report describes the preliminary design effort for a very wideband high power tracking radar. Consideration has been given to dispersion effects of radar components, phase and amplitude tolerances required to meet specified range sidelobe levels, availability of components, etc. An approach for modifying an existing radar is also presented.

TABLE OF CONTENTS

Section	Title	Page
I	OBJECTIVES	1
II	INTRODUCTION	3
III	APPROACH	5
	A. General	5
	B. Present System Description - Floyd Test Annex	7
	C. Existing Equipment	7
IV	NEW EQUIPMENT REQUIREMENTS	13
	A. General	13
	B. Signal Processor Subsystem	13
	C. Feed Horn and Receiver Subsystem	16
	D. Transmitter Subsystem	17
	E. Data Handling Subsystem	18
V	DESIGN CONSIDERATIONS	21
	A. Antenna and Feed Horn	21
	B. Transmitter Design Criteria	32
	C. High Voltage Supply and Cooling System	35
	D. Receiver Design Criteria	37
	E. Signal Processor	39
	F. Data Handling	49
	G. Servo System	57
VI	MODIFICATION PROCEDURE	67
VII	SUMMARY	69
VIII	APPENDIX A: System Sensitivity Calculations	75
	APPENDIX B: Acquisition and Tracking	79
	APPENDIX C: Distortion Analysis of Broadband Networks	85
	APPENDIX D: Power Amplifier	103
	APPENDIX E: System Specifications	107

LIST OF ILLUSTRATIONS

Figure	Title	Page
Frontispiece	Floyd Test Annex	
1	Phase I Radar Parameters	6
2	60-Foot Cassegrain Antenna	9
3	Orbital Computer	11
4	Control Console	12
5	Equipment Room	14
6	Operations Building	15
7	Modes of Operation	16
8	Signal Processor Block Diagram	17
9	Feed Horn Design	18
10	Receiver System Block Diagram	19
11	Interim Transmit System	19
12	Data Handling System	20
13	Cassegrain Antenna Principle	22
14	Parameters of Cassegrain Antenna	23
15	Aperture Illumination Function	24
16	Aperture Blockage	24
17	Pattern Variation with Frequency	27
18	Five-Horn Feed Schematic	29
19	Monopulse Sum & Difference Patterns	31
20	Simplified Basic Design	40
21	Amplitude and Phase Errors	41
22	Effects of Non-Linearities	43
23	Transmitter Phase Response	46
24	Hydraulic System Performance	61
25	Typical Hydraulic Valve Characteristics	62
26	Servo System Transfer Response	63
27	System Parameters	72

Floyd Test Annex

SECTION I
OBJECTIVES

The objective of this program is to construct a very wide band radar to act as an experimental test instrument to do research in satellite object identification techniques. In addition to this, the radar will also be used to develop and test wide band radar calibration techniques, wide band performance measurement techniques, wide band signal processing concepts and as a test bed to test and evaluate wide band radar components. A number of these wide band concepts are under development as part of the exploratory research program under Project 4506. The radar is being designed such that future extension to a base line configuration will be possible with no major redesign.

SECTION II
INTRODUCTION

For a number of years, . research requirement has existed for a very high range resolution sensor of sufficient sensitivity to perform exploratory research in a number of problem areas in object identification. Extensive laboratory research and development has already been successfully accomplished in the areas of very wide band pulse compression signal generation and processing required to successfully implement such a facility. Implementation of these techniques into a practical system gives rise to a number of additional problems which must be resolved. Developing the specific waveform generators and signal processing technique to achieve specified sidelobe levels is therefore only part of the problem. In practice, the dispersion effects of all the components in the radar system exert an effect upon the overall performance. This becomes a particularly difficult design problem in a very wide band radar where stringent phase and amplitude linearity requirements must be met. In order to adequately examine the practical problem involved, it is necessary to construct and evaluate an experimental model of such a radar. This report describes the approach taken in the design of such a radar which makes maximum use of available components and facilities and represents the best trade-off in cost and performance.

During the course of this study, a number of problem areas concerning the design and implementation of a very wide band radar were uncovered and are described herein. Further study of these problem areas will be accomplished during the development phase of the equipment.

SECTION III
APPROACH

A. GENERAL

As a result of the preliminary design study, it became evident that if excessive losses and dispersive effects were to be avoided, all of the wide band transmitting and receiving components must be located as close as possible to the feed structure avoiding the use of rotary joints. This approach permits the use of minimal length waveguide runs to minimize dispersion and loss effects and the elimination of potentially serious "wow" and dispersion which a rotary joint would almost certainly generate. Furthermore, bandwidth reduction should be accomplished as early as possible in the receiver chain in order to minimize dispersion and/or phase shift effects which would most likely be encountered if transmission of the wide band signal over extended distances were required. Although many pulse compression techniques could meet this requirement, certain techniques were more desirable than others because a minimum of pedestal mounted equipment would be required, avoiding weight, inertia, maintenance and overhaul difficulties. Signal processing and data handling equipment would require extremely delicate adjustment, careful monitoring and other attention, and thus should be located in the main building free from vibration under temperature and humidity control and should allow free access to those operating and maintaining it. This will normally require fairly long reduced bandwidth cable runs between the wide band receiver components mounted on the moving portions of the antenna and these equipments. Here again, dispersion and phase shift effects must be avoided, although the problem is far less serious at the reduced bandwidth. A cable wrap appears to be the most logical means of interfacing the moving platform with the fixed pedestal in order to avoid the potentially serious dispersion and phase shift effects inherent in most rotary joints, and the cross talk and limited bandwidth inherent in slip rings.

In searching for a suitable location for the proposed sensor, existing RADC Test Sites (among them the Passive Satellite Research Terminal, (PRST) located at Floyd, N.Y.) were examined. The site should be convenient for daily travel by personnel, should have most of the necessary support facilities already available (power, water, buildings, etc.) and should have an acceptable field of view for objects of interest. The PRST fulfilled these requirements and, in addition, due to termination of project activity, a complete sensor was available. The design effort was then concentrated on the possibility of modifying this facility.

Apart from the practical considerations of choosing sites and apparatus capable of being modified, the overriding system design considerations were tied to more elegant requirements. These included obtaining adequate transmitter power and bandwidth at a carrier frequency of 3350 MHz, and configuring a waveform generator and signal processor to perform the desired measurements with adequate fidelity, with a further bandwidth reduction to allow realizable A/D conversion and recording.

The center frequency of 3350 MHz has been assigned as the operating frequency for the system. The choice of signal bandwidth and transmitted power are synonymous with selecting a tube which has permissible levels of phase and amplitude distortion, while providing a reasonable average power and instantaneous bandwidth. The search for a tube ultimately narrowed down to the VA145 hybrid traveling wave tube which had an advertised average power on the order of 10 kw, a bandwidth up to 300 MHz with apparently acceptable levels of phase and amplitude distortion.

A pulse repetition frequency of 70 pulses per second was chosen to be consistent with maximum average power capabilities of the chosen tube at its maximum reliable pulse length, while providing an acceptable unambiguous range. The choice of pulse repetition frequency was based upon a number of factors including unambiguous range, doppler and average power.

Since a maximum bandwidth of 300 MHz could be expected from the VA145 tube with a reliable pulse length of 20 microseconds, the decision was made to develop an interim Phase I radar using this tube with a 20 microsecond pulse width and a 250 MHz bandwidth.

A Phase II capability of 500 MHz bandwidth and a 40 microsecond pulse length were selected as the parameters for the final system, since a 500 MHz bandwidth results in a nominal range resolution of one foot. The 40 microsecond pulse length results in a 20 kw average power capability at the same prf. The system detection capability was calculated for a 0.1 square meter target which was considered a reasonable estimate of the size reflector which might be encountered with range resolution near one foot. The anticipated performance in Phase II would provide a 13 db S/N for a 0.1 square meter target at 600 n miles. This range performance is adequate for a test facility radar sensor. In order to minimize weak signal degradation when observing complex targets it is desirable to keep the waveform temporal sidelobes as low as possible. A time sidelobe specification of -35 db was selected from target dynamic range considerations and expected state-of-the-art limitations for a system with a time bandwidth product of 20,000.

Since the transmitter provides the major limitation to the attainment of 500 MHz instantaneous bandwidth, the feed horn assembly, RF components, receiver front ends and data handling equipment were configured to perform at the full bandwidth and power, thereby eliminating the need for their modification when the system is upgraded to Phase II performance. The waveform generator and signal processor require relatively minor modification to make the transition from Phase I to Phase II operative. The Phase II modification then will consist primarily of replacing the output tube with a full bandwidth higher power tube with an attendant modification of the pulse forming network and a minor increase in the waveform generator and signal processor capability. The change to the signal processor is not in any way as complex or difficult as the increase in time-bandwidth product from 5000 to 20,000 might indicate. Figure 1 is a tabulation of Phase I parameters used in subsequent calculations.

P_{AV} : 10 KW	BANDWIDTH: 250 MHz
PRF: 70 PPS	λ : 0.1 METER
PULSE WIDTH: 20 μSEC	σ: 0.1 SQ. METER
G_t : +50 db	LOSSES: -5 db
G_r : +50 db	NF +6 db
BEAMWIDTH: 0.3 DEG.	FREQUENCY 3350 MHz

POLARIZATION: TRANSMIT: VERTICAL
RECEIVE: VERTICAL & HORIZONTAL

Figure 1. Phase I Radar Parameters

B. PRESENT SYSTEM DESCRIPTION, FLOYD TEST ANNEX

The present passive satellite research terminal is located at the Floyd Test Annex approximately seven miles from Griffiss AFB. It was originally configured as a large ground-based radio facility designed primarily to demonstrate point-to-point communication over long distances via reflections from satellites such as Echo [1]. The station consists of a 60-foot steerable paraboloidal reflector. The accuracy of the reflector surface is such that full aperture efficiencies will be achieved at frequencies as high as 10 GHz. Tracking accuracies of 0.02 degrees are achieved through the use of a precise hydraulic servo and drive system. The system also contains high power communication, transmitters, low noise parametric amplifiers, frequency synthesizers and an orbital programmer.

In addition to the antenna and pedestal, the facility includes an optical tracker; a heat exchanger building; an operations building containing the operating console, equipment room and high voltage power supply; a microwave relay link; a mobile service tower; an emergency power generating system and a boresight tower located approximately 2.9 miles from the antenna.

C. EXISTING EQUIPMENT

A detailed survey was conducted to determine the applicability of the present equipment and the extent of the modifications required to convert the system to a high resolution tracking radar. The present facility includes the AN/FRC-40, consisting of the following major components:

- 60-foot Cassegrain antenna
- Pedestal, including antenna drive and servo system
- Feed Horn - dual polarized S-band monopulse feed with a central X-band dual polarized feed
- Angle track computer
- Control console
- S- and X-band receive systems
- Noise figure measurement system
- S-band transmitter
- 20-kw power supply and heat exchanger
- Frequency synthesizer
- Time code generator ASTRO Data 6190-133
- Frequency standard HP-107BR
- Optical tracker

A new power supply, heat exchanger and transmitter are presently being installed to implement the X-band system. This power supply is a copy of the supply for the Haystack radar X-band transmitter and is capable of providing 120 kv at 15 amp or 40 kv at 60 amp.

The present S-band system operates in the 1700-2400 MHz band and has been used to track Echo I and Echo II to conduct experiments in satellite communications. The new frequency authorization will be centered at 3350 MHz with a ± 250 MHz bandwidth.

Primary considerations affecting the feasibility study were the emphasis on maximum utilization of existing equipment without performance compromise and accomplishing a major portion of the system design in-house. The site survey performed during the feasibility study resulted in a wealth of data which is summarized in the following paragraphs.

Antenna. The Cassegrain reflector system is usable without modifications. Due to the change in frequency, however, the feed horn must be replaced. An in-house study was initiated to examine various feed horn configurations and to develop specifications and cost data. The results of this study are presented later in this report. The present antenna beamwidth of 0.5° will become 0.3° with the conversion to a higher frequency. Figure 2 is a view of the 60-foot diameter antenna.

Pedestal and Servo System. No changes to this portion of the equipment are contemplated until a study is completed on the actual servo performance. The servo system was designed to have an average velocity rate of two degrees per second and an acceleration rate of one degree per second per second.[2] The servo system analysis is presented later in this report. The antenna drive is composed of the azimuth bull gear and the elevation sector gear and hydraulic motors. The antenna may be driven 720 degrees in azimuth, being limited by cable windup on the maypole, and 90 degrees in elevation. Two hydraulic motors, connected to eliminate backlash, are used for each drive. The servo system is designed to achieve mean tracking accuracy of 0.03 degrees in the absence of wind. An electro-hydraulic servo valve is used to control the hydraulic drive motors. The servo will accept both analog and digital commands. Digital shaft encoders are mounted on the antenna to provide position data readouts for both azimuth and elevation to an 18 bit accuracy. Analog error sensors permit control of the antenna by a remote optical tracker as well as an analog scan generator to program a desired scan pattern.

Angle Track Computer. The angle track computer consists of the following equipment:

- Packard-Bell FX-1 Flexowriter.
- Packard-Bell PB-250 Digital Computer.
- Packard-Bell HSB-22 Buffer.
- Packard-Bell MX-1 Memory.
- Packard-Bell HSR1 High Speed Reader.
- Ampex MTU-1 Tape Handling Unit.
- Ampex FR-400 Tape Recorder.

Further investigation is necessary to determine whether modifications are required. Computer programs already exist for tracking satellites of the Echo class.[3,4] New programs have been written to track lower orbiting satellites.[5] The present digital sampling rate may be inadequate to provide the overall servo bandwidth required to adequately track these lower orbiting satellites. Possible redesign may be required as determined by studies.

The PB-250 digital computer, in conjunction with the magnetic tape unit, provides general purpose computing and terminal data recording functions. The magnetic tape unit is an input/output device as well as an auxiliary memory for the computer. Since the computer uses a serial memory, a buffer is required between the computer and external equipment. This buffer is a 22-bit shift register and performs a one-word serial-to-parallel conversion. A high speed photo-reader and a flexowriter are used for computer control and data printout. Station keeping functions such as antenna position, doppler shift, terminal operating conditions, and other miscellaneous functions are also recorded by the digital computer in blank spaces provided on the magnetic tape.

During target acquisition, the program presents the "most probable" satellite ephemeris to the antenna servo channels. The scan is added and the composite signal controls the antenna. After lock-on has been accomplished, the program utilizes actual track position of the satellite to determine a set of interpolation constants for use in programmed tracking.

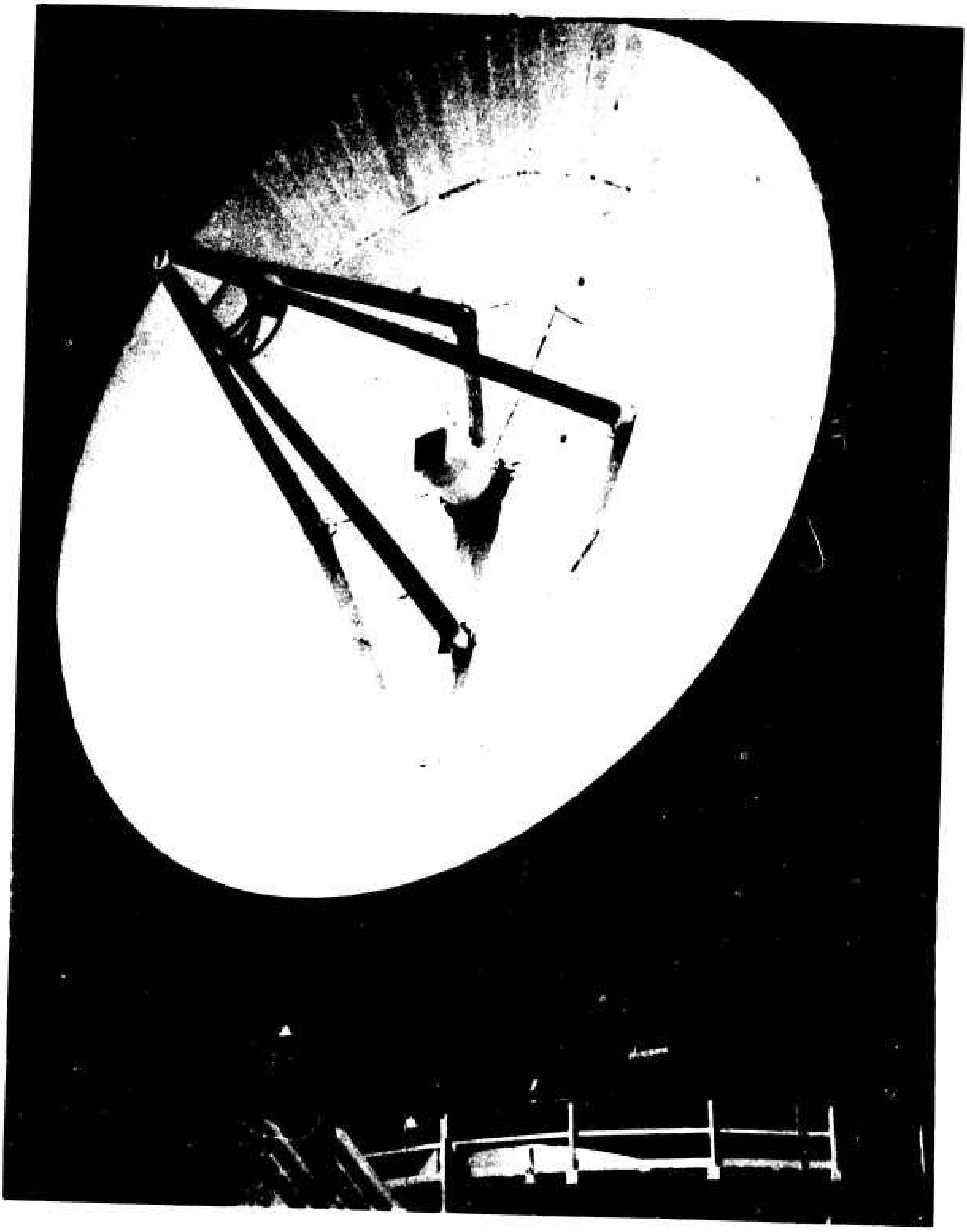


Figure 2. 60-Foot Cassegrain Antenna

In programmed tracking, the antenna is directed by the program which interpolates between a series of pre-computed probable orbits. These pre-computed orbits are stored on magnetic tape and correspond to variations in expected time-of-arrival from the "on-time" orbit. Figure 3 is a view of the PB-250 orbital computer.

Control Console. Minor modification to the control console will be required to add remote control for new equipment and other added functions. Figure 4 is a view of the present control console.

The control console contains data readouts and manual controls for operation of the terminal. The control console permits the operator to interpret the output of the electronic and electro-mechanical equipment and to respond by introducing the commands required to accomplish the mission. The antenna control displays the azimuth and elevation position of the antenna by a pair of four-digit decimal readout devices and also indicates the degree of maypole cable windup. Manual control of the antenna is accomplished at this control console. The console permits the operator to control the modes and limits of automatic sequences between acquisition and other track functions. The transmitter control consists of status, voltage, power, on-off displays and controls for the operation of the transmitter. The status of safety interlocks and alarms which indicate failure is displayed on the control console.

Receive System. The present communications receive system is unusable in the new configuration. A completely new receive system will be designed to replace it. The angle track receiver will be usable with minor modifications to the IF portions to accommodate the bandwidth of the pulsed radar. New angle track receiver preamps will have to be designed.

Transmit System. The present final amplifier, power supply and heat exchanger cannot be used in the new configuration. A new transmit system will be designed using the new power supply and heat exchanger designed for the X-band communications system. This new power supply and heat exchanger are described later in this report.

Station Clock. The station clock consists of an ultra-stable oscillator, HP-107BR, a time code generator Astro Data 6190-133, a WWV receiver and time displays. The ultra-stable oscillator is the station frequency standard which supplied 5 MHz, 1 MHz and 0.1 MHz signals to the time code generator and frequency synthesizers. The stability of all signals is better than one part in 5 x 10^{10} per day. The time code generator supplies time-of-day in hours, minutes and seconds to the time displays on the control console and the station clock. Binary serial time-of-day and timing pulses are routed to the digital computer. The WWV receiver is used as a calibration standard for the stable oscillator. An auxiliary oscilloscope, supplied with one-second timing marks, is used to permit time comparisons against signals received on the WWV receiver to establish a frequency drift record for a frequency standard.

Frequency Synthesizer. The present frequency synthesizer cannot be used in the modified radar without major redesign since the output frequencies were determined for the communications operation. A new frequency synthesizer will be incorporated into the signal processor design and synchronized by the station clock.

Auxiliary Equipment. Other auxiliary equipment presently installed at the site includes a two-ton air conditioner and a 15 KW heater mounted on the antenna back-up structure adjacent to the center room. The purpose is to maintain the temperature of the antenna mounted RF equipment rooms at 70 $\pm$ 2 degrees F.

A 20-ton air conditioner cools the main equipment room in the operations building. Cool air is ducted directly into the room and warm air is drawn off the equipment.

A diesel-powered generator, capable of turning on 30 seconds after power failure, provides power for interior lights, radome blowers and stowage of the antenna. A battery is used to operate station clock. Also available are compressed air facilities as well as a mobile service tower (Hi Ranger) for access to and maintenance of the antenna mounted equipment.

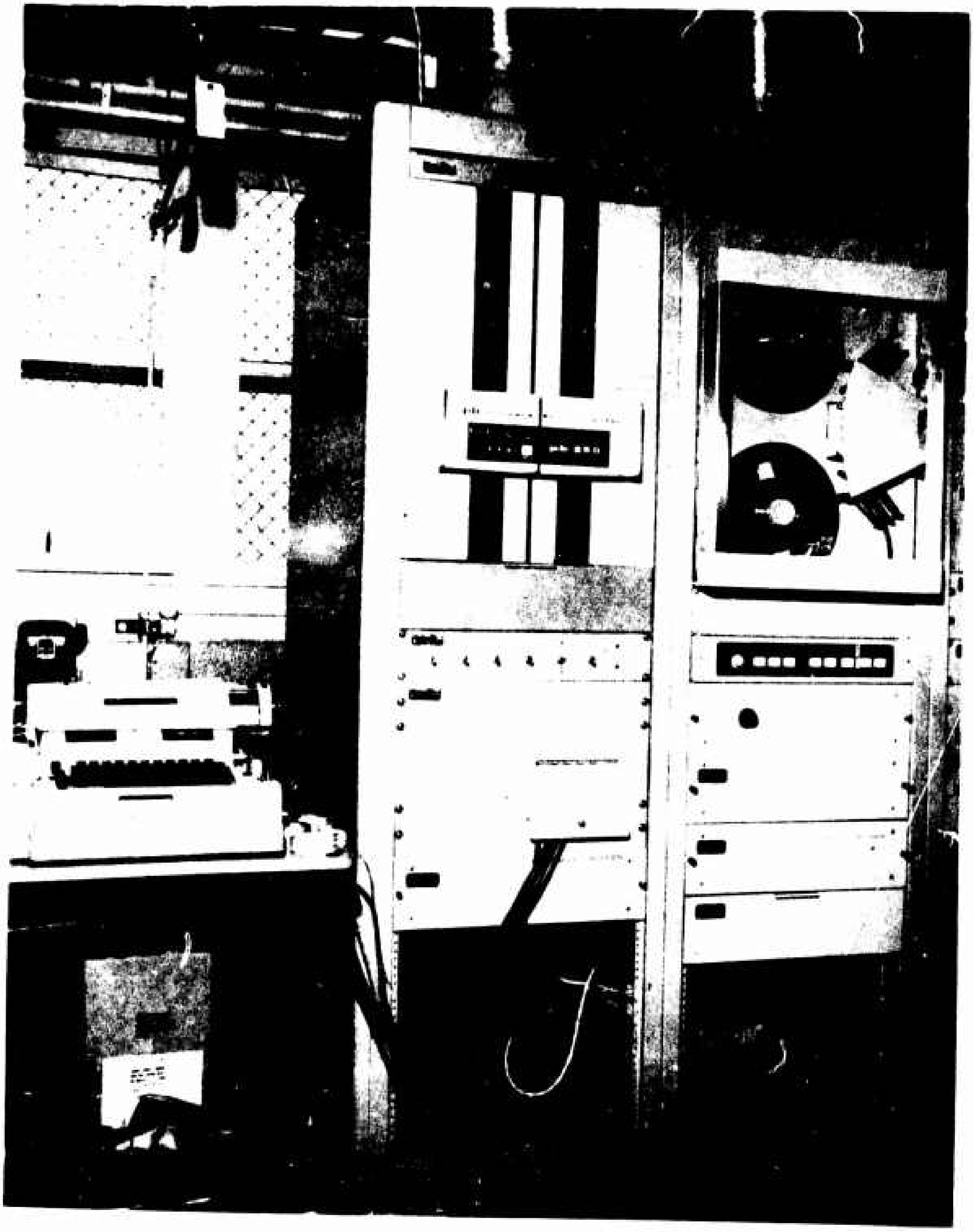


Figure 3. Orbital Computer

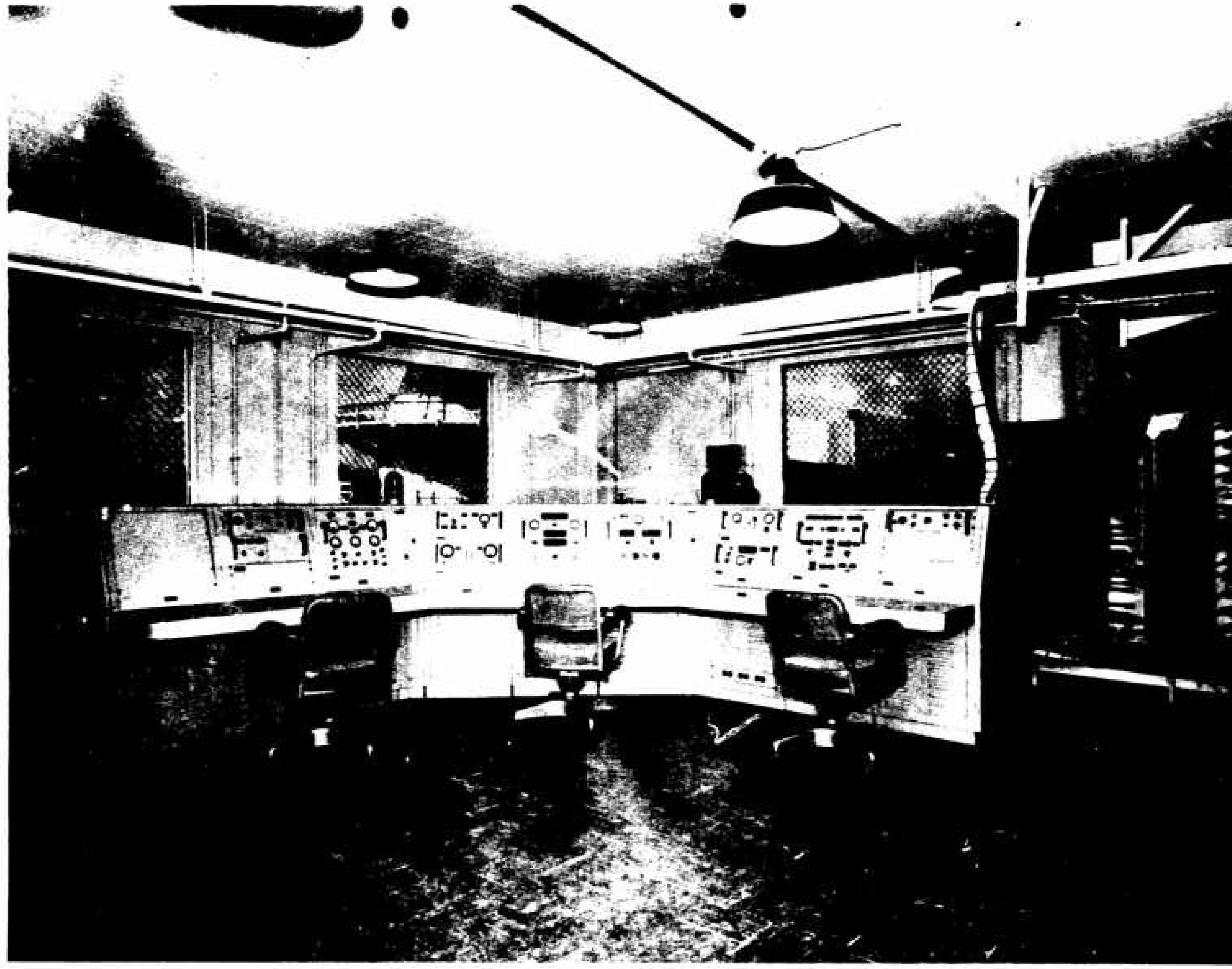


Figure 4. Control Console

SECTION IV
NEW EQUIPMENT REQUIREMENTS

A. GENERAL

In addition to the items described previously, new equipment must be designed and procured. This consists of a new feed horn with its associated microwave plumbing; a new transmitter and modulator; a new receiver front end including duplexers, etc.; a signal processor which includes the waveform generator, transverse equalizer, phase and amplitude detectors, and precision range tracking circuits; data handling and recording equipment.

Minor modifications to some existing hardware will be necessary. The present noise figure measurement system will also be modified and incorporated into the new receiver subsystem. Studies will continue to be performed on upgrading the servo system performance[6] and the orbital computer sampling rate.

All wideband portions of the system subject to dispersion effects will be located on the antenna pedestal. These include the transmitter, receiver, duplexer and other microwave hardware, noise figure measurement system and the wideband portions of the signal processor. All other portions of the system will be placed in the equipment room shown in Figure 5 located at the rear of the control room. These include the orbital computer, signal processor, data handling, servo cabinets, range tracker, station clock, etc. Figure 6 is a plan view of the operations building.

B. SIGNAL PROCESSOR SUBSYSTEM

The signal processor includes the waveform generation and compression, range and angle tracking function, generation of system timing signals, local oscillator and coherent references for the entire system and provides distortion correction for the entire system to ensure that the range side lobe level of -35 db is achieved.

The timing and synchronizing signals for the radar are derived by the signal processing subsystem from the system PRF generator and the station clock to ensure that the system remains coherent throughout and stable with respect to a time reference.

The radar will operate in three basic modes as shown in Figure 7. The purpose of multiple mode capability is to ensure a smooth transition from the CW acquisition to the high resolution mode. Mode I will consist of transmitted waveform with no frequency modulation. It will have a duration of 20 microseconds in Phase I and 40 microseconds in Phase II. The Mode I operation will be used for initial acquisition in range and angle using active monopulse angle tracking in two dimensions.

Once the object of interest has been acquired and sufficient angle track data has been acquired to permit the computer driven angle track system to take over, a transition will be made into Mode II. The Mode II operation utilizes a simple linear FM waveform of 2.5 MHz bandwidth. Its duration is 20 microseconds in Phase I or 40 microseconds for Phase II. The intent of the Mode II operation is to improve the system range resolution to a nominal 200 feet and initiate active range track preparatory to entering the high resolution mode or Mode III. The operation in Mode II will derive range-rate data from the target to feed the third order range tracking system required in the high resolution mode since the transmitted pulses are far too short to result in accurate doppler measurements. No active angle tracking will occur in Modes II or III.

Figure 5. Equipment Room

Mode III will consist of a linear FM transmit waveform of 250 MHz bandwidth in Phase I and a 500 MHz bandwidth for Phase II. The resolution will be nominally two feet and one foot, respectively, however, meaningful measurements will be taken with a weighted waveform resulting in range resolution of four feet and two feet, respectively, for adjacent targets differing in amplitude by 34 db.

The signal waveform is obtained by combining the 2.5 MHz Mode II ramp with either a 247.5 MHz or a 497.5 MHz ramp for actual transmission of 250 MHz or 500 MHz signals. Included in the signal waveform is a round trip transmission path predistortion function with a correction for atmospheric dispersion in the 500 MHz case.

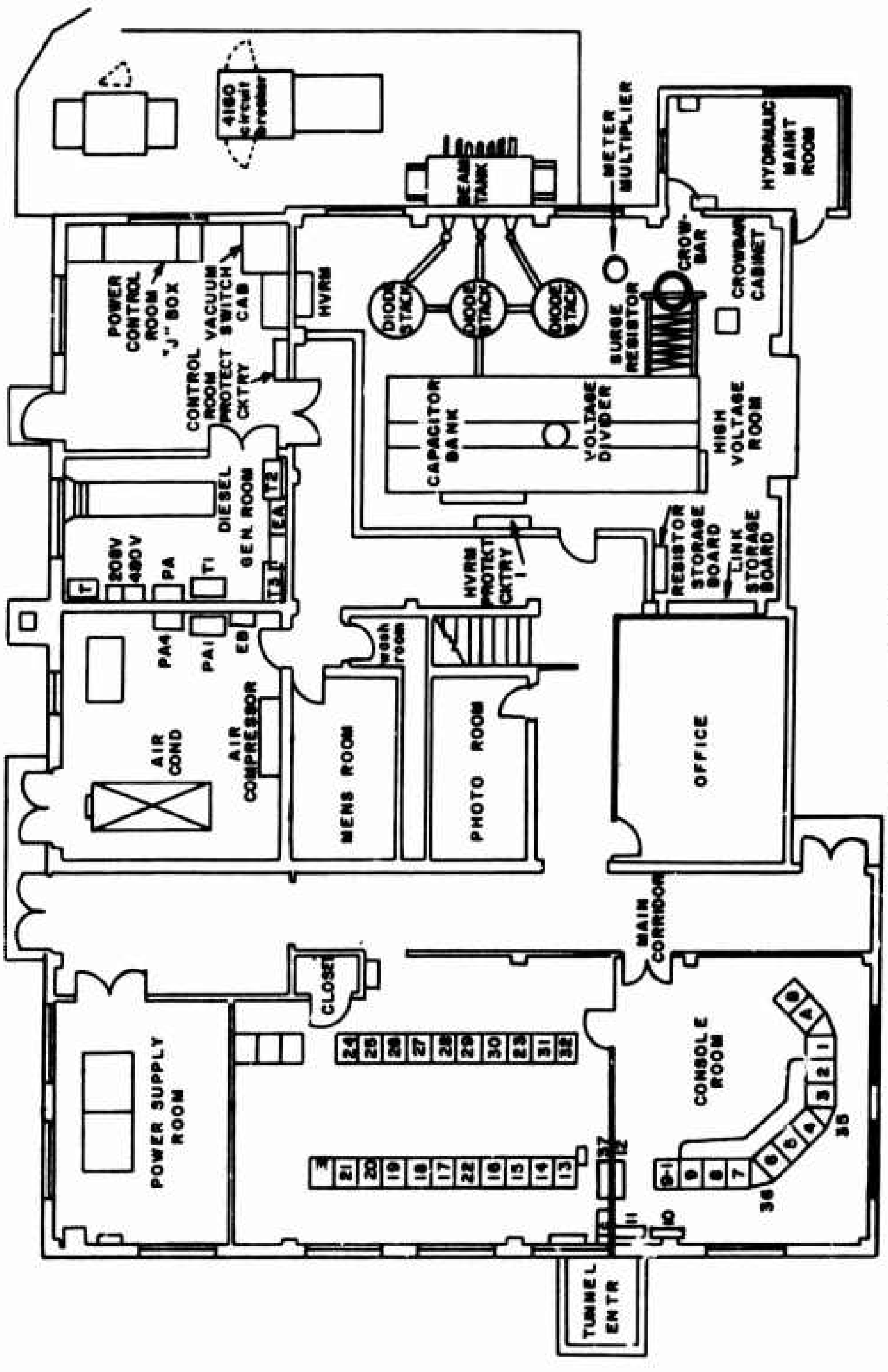


Figure 6. Operations Building

	PHASE I		PHASE II	
	RANGE RESOLUTION	TIME-BW PRODUCT	RANGE RESOLUTION	TIME-BW PRODUCT
MODE I (ACQUISITION)	2 MILES	1	4 MILES	1
MODE II (2.5 MHz LIN. FM)	200 FEET	50	200 FEET	100
MODE III (MEASUREMENT)	2 FEET	5000	1 FOOT	20,000

Figure 7. Modes of Operation

When the signals are received, they are mixed again with a frequency off-set ramp pre-corrected for target radial velocity resulting in IF signals containing a 2.5 MHz linear FM modulation with a duration of either 20 microseconds or 40 microseconds. The range resolution of the full transmitted bandwidth is retained by virtue of the fact that the IF signals have a center frequency which varies with range at the rate of 25 kHz per foot. This frequency difference is then transformed into range resolution in the time domain by passing it through an inverse linear FM compression filter which has a time delay sensitivity of 125 kHz per microsecond in Phase I and 62.5 kHz per microsecond in Phase II. These signals when displayed within the 20-microsecond window (40-microsecond in Phase II) provide range resolution of two feet across a 100 foot aperture in Phase I and range resolution of one foot across a 100-foot aperture in Phase II. In either case, the output IF bandwidth is 2.5 MHz, allowing a feasible A/D conversion and recording system to be employed.

The received signals will be weighted at IF after compression in a transversal equalizer to provide the required side lobe level.

An interesting and attractive aspect of this system configuration is that only the wideband portions of the signal processor need to be located on the antenna structure and all of the narrow band portions, including the pulse compression networks, range tracker and angle tracker will be located in the operations building.

The actual output information from the Signal Processing Subsystem will be either amplitude and phase or I and Q signals derived for the duration of the 100-foot range window. The radar will be making the high resolution measurements on two simultaneously received orthogonal polarizations by delaying the signal in one polarization channel and processing both signals through the same circuits to alleviate channel matching problems.

A mode sequencer is included in the signal processor to sequence the system from Modes I through II but also includes the capability of interlacing angle track pulses between the Mode III signals in the event that computer angle tracking is inoperable. A simplified block diagram of the signal processor is shown in Figure 8.

C. FEED HORN AND RECEIVER SUBSYSTEM

The system will transmit vertical polarization only but will receive both vertical and horizontal polarization in the wideband channels. For initial acquisition and angle tracking in Mode I, narrow band monopulse configured vertically polarized receive horns will be used. This results in a feed horn design as shown in the simplified drawing of Figure 9.

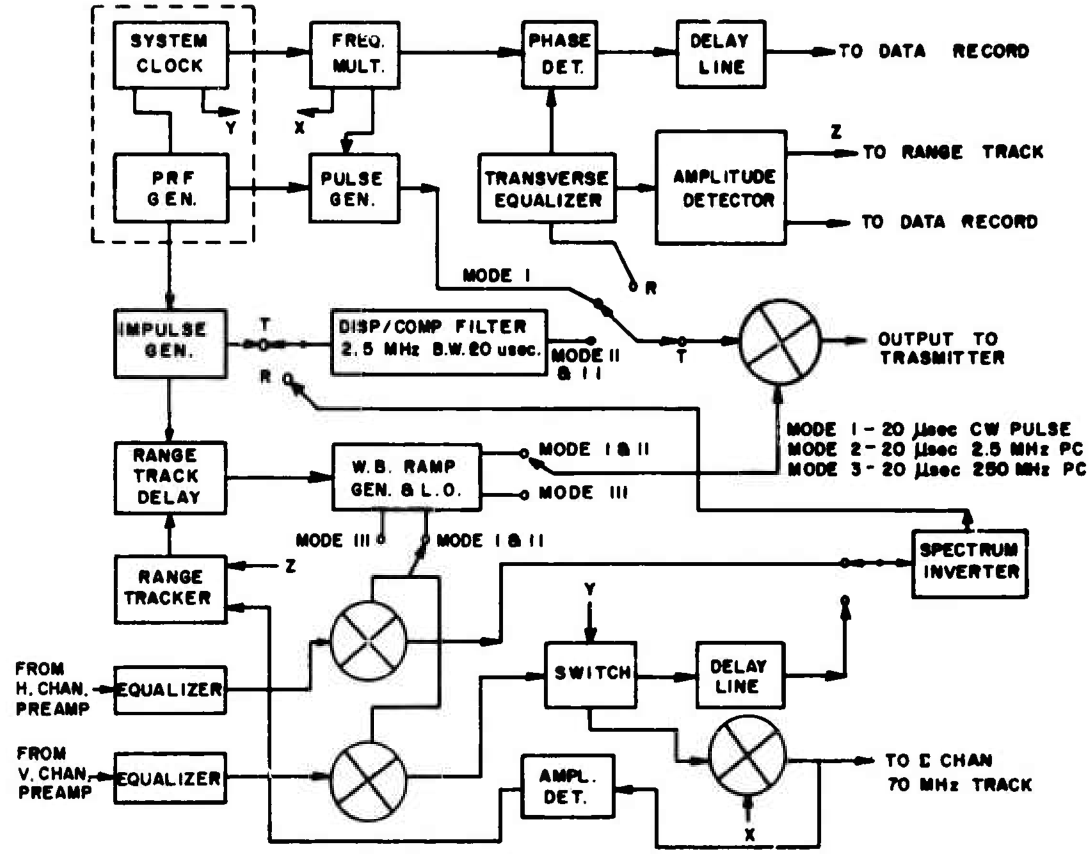


Figure 8. Signal Processing Block Diagram

Wideband tunnel diode amplifiers will be used in the information channels. The two channels will be equalized so that the relative phase and amplitude of the received information will be preserved. The wideband receiver will be designed for the full 500 MHz bandwidth capability so that no modifications to the receiver will be required in Phase II. The significant parameters of the receive subsystem are the phase and amplitude tolerance which must be held to insure no degradation of the signal processor side lobe level performance. Figure 10 is a block diagram of the receiver subsystem. The output of the wideband receiver will be fed to the signal processing system for waveform compression and detection. A survey of existing receiver technology indicates that tunnel diode amplifiers are feasible for this application.

The narrow band sum and difference channels will also be tunnel diode amplifiers. These channels will feed the angle track receivers in the active track mode.

D. TRANSMITTER SUBSYSTEM

The wideband transmitter will be designed to operate from the existing 1.2 MW high voltage supply and heat exchanger. The transmitter will be implemented initially with an available interim tube modified for higher frequency operation and will be constructed so that minimum modifications will be required for conversion to Phase II operation. It is expected that, with proper design, only the pulse forming network will have to be replaced to

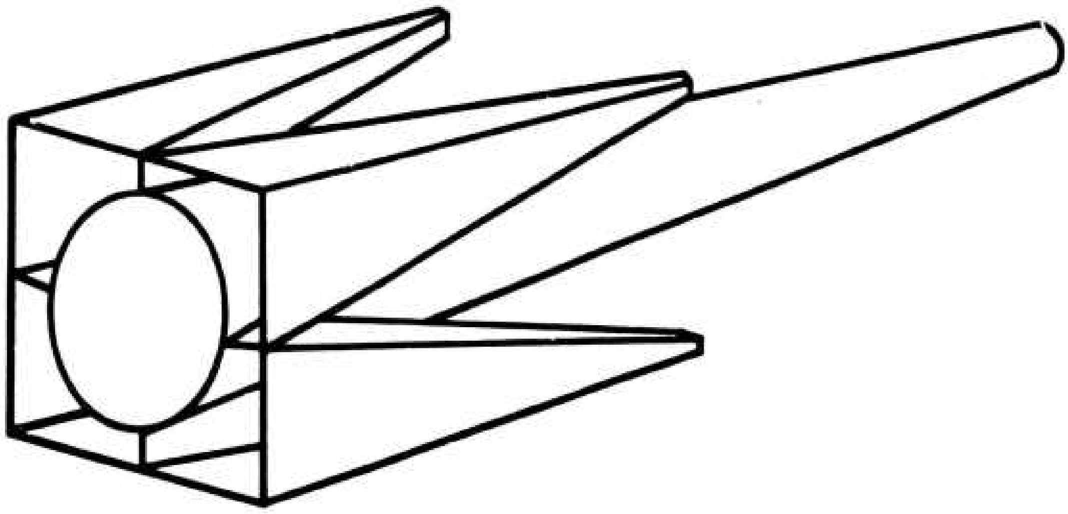


Figure 9. Feed Horn Design

accommodate the new pulse width in Phase II. The Phase II tube will be designed to use the same tube socket as the interim tube. A new focusing solenoid will also be supplied with the Phase II tube. Pulse leveling will be included in the modulator to minimize pulse ripple. Phase monitoring and correction circuits will be included in the driver section of the transmitter to maintain the specified linear phase vs frequency characteristic over the signal bandwidth.

The interim tube will be a Varian Hybrid TWT (VA145F). Band center will be at 3350 MHz. The tube will have an instantaneous 3 db bandwidth of 250 MHz. Peak power will be between four and six megawatts with a 0.0028 duty and a 20 microsecond pulse. The traveling wave section of the tube is of the coupled cavity type.

The Phase II tube will be designated the VA915A. It will have a 500 MHz electronic bandwidth at the 6-db points and will be capable of 40-microsecond pulses. Cathode pulsing will be employed. Peak power over the band will not exceed 10 megawatts and the average power when employing a linear FM ramp will be twenty kilowatts. The type of circuit to be employed in the TWT section has not been firmly established. It will probably be of the centipede type. However, if unexpected problems are encountered, the coupled cavity circuit may be used. The achievement of a tube with a high degree of phase linearity and stability is extremely important and will receive considerable effort in the tube's design.

A block diagram of the transmitter subsystem is shown in Figure 11.

E. DATA HANDLING SUBSYSTEM

The output of the signal processor is then encoded in a pair of 8-bit 10-MHz A/D converters. A logarithmic amplifier will be used to maintain dynamic range of the input signal. The encoded information is then stored in a buffer memory along with other data such as timing signals, identity, range, azimuth and elevation angles and doppler correction. The information is recorded on 16-channel tape operating at 120 inches per second. At the completion of a data run the tapes are played back at a five to one speed reduction and reformatted on seven-channel tape in binary form for later computer analysis. Figure 12 is the block diagram of the Data Handling System.

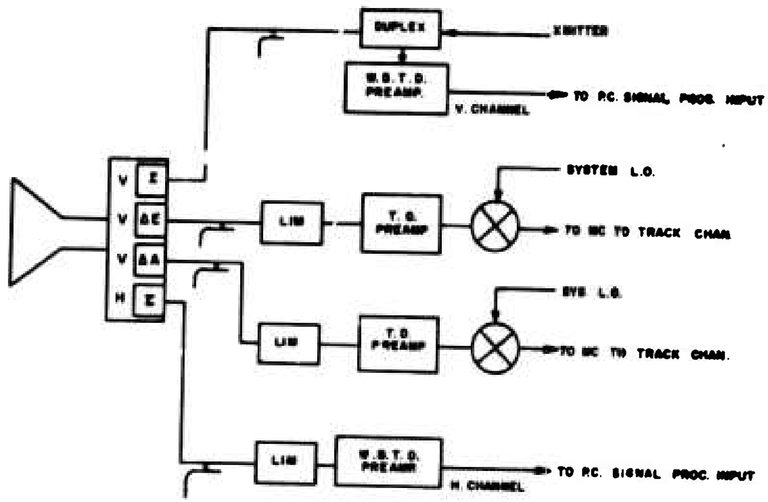


Figure 10. Receiver System Block Diagram

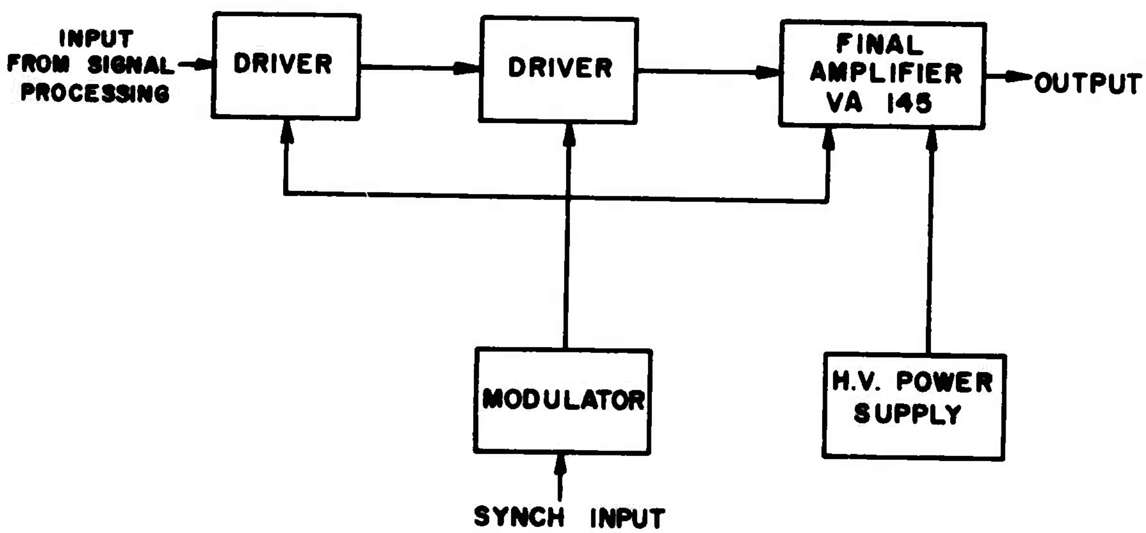


Figure 11. Interim Transmit System

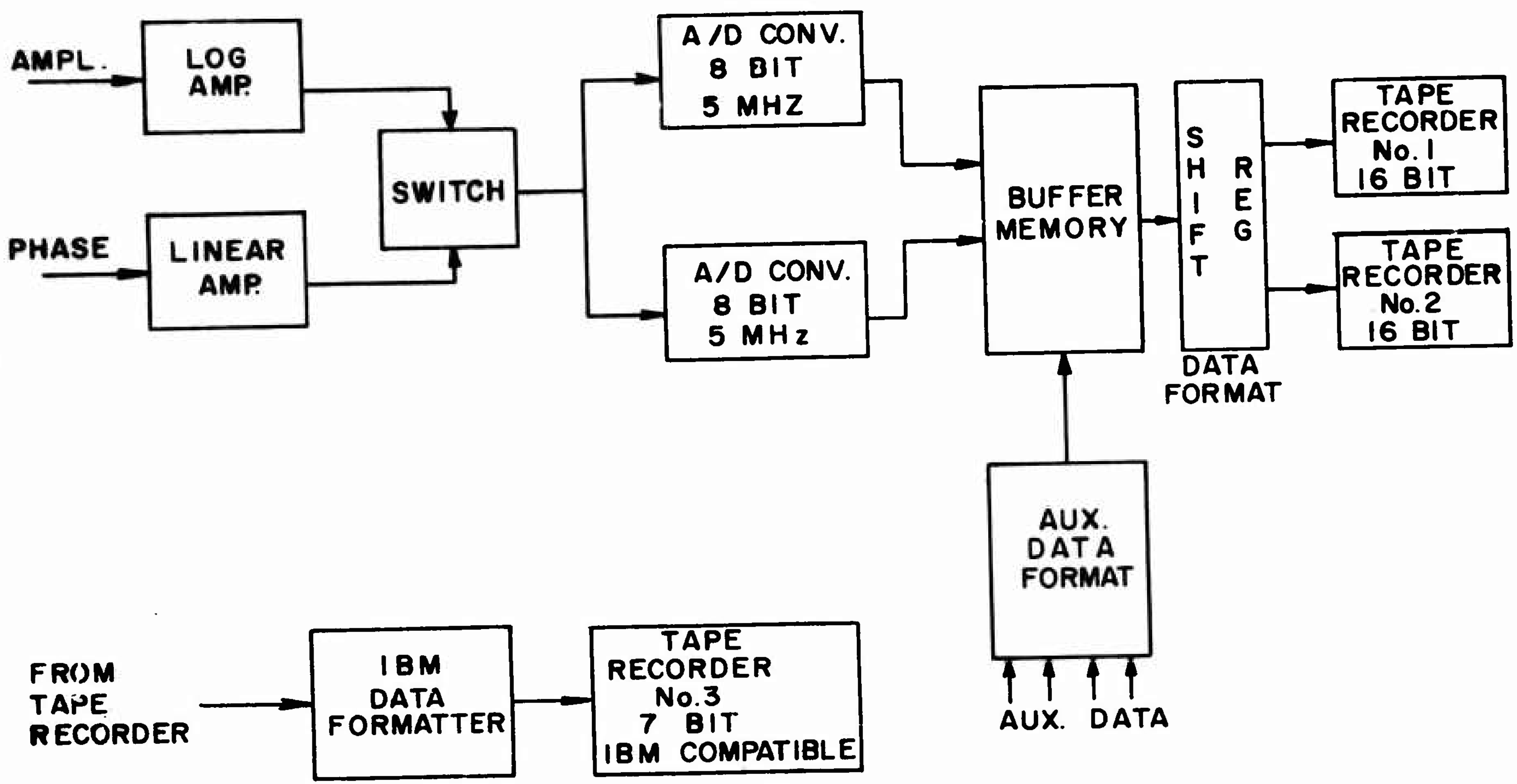


Figure 12. Data Handling System

SECTION V
DESIGN CONSIDERATIONS

A. ANTENNA AND FEED HORN

This analysis concerns the preliminary antenna design considerations required to transform the present Satellite Communication Terminal AN/FRC-40 into a Signal Processing Test Facility. The modified radar is to be a high power, broad bandwidth, high performance system capable of tracking orbiting space objects to obtain high resolution amplitude and phase profiles.

As a result of a preliminary evaluation of the present AN/FRC-40 antenna, it was concluded that the existing antenna design could be implemented for the radar. Furthermore, the only major modification required for implementation is a new feed capable of operating over the required 500-MHz bandwidth. Since this broad bandwidth capability only applied to the data gathering function of the radar, the feed can be designed to operate in an independent tracking mode at a reduced bandwidth (150 MHz). Functionally, the feed will transmit broadband vertically polarized energy only, but will receive both vertically and horizontally polarized signals. These signals will be transmitted through identical channels to separate broadband receivers for signal processing and data gathering. Target acquisition and tracking will be accomplished using only the vertically polarized return signal in narrow band tracking channels. The two radar functions, i.e., data gathering and target tracking, will be performed by time sharing the transmitted pulse train in a prespecified order.

Contained herein is a preliminary evaluation of the existing antenna establishing its parameters under the operational environment. Also included is a discussion of the design consideration on the new feed to convert the AN/FRC-40 antenna for use in the modified radar.

1. Analysis of AN/FRC-40 Antenna

a. Present Antenna Design

The AN/FRC-40 antenna is a dual reflector configuration of the Cassegrain design. The principle of the Cassegrain antenna is shown in Figure 13. The feed is located on the system axis at a point between the hyperbolic subreflector and the parabolic primary reflector. The feed generates a spherical wave which illuminates the subreflector. The subreflector in turn illuminates the paraboloid which reradiates a collimated beam in the forward direction. The exact locations of the hyperboloid and feed are determined by the equations of the surfaces and the system geometry. The phase center of the feed is located at the real focal point of the hyperboloid while the virtual focus of the hyperboloid is coincident with the focus of the parabolic surface.

The Cassegrain design offers some distinct advantages which make it attractive for use in the modified radar. These include:

- Compression of the axial length of the antenna system.
- Elimination of long transmission lines.
- More flexibility in primary feed design than is possible with front fed paraboloids.

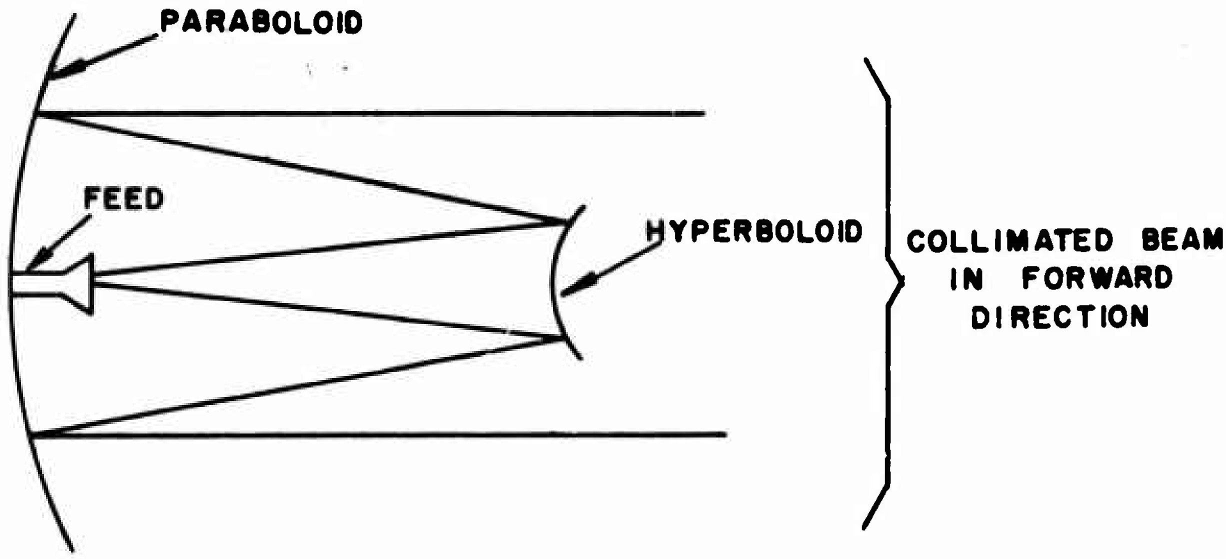


Figure 13. Cassegrain Antenna Principle

- Provision of another degree of control in illuminating the primary reflector by using an auxiliary reflector.

b. Specifications of the AN/FRC-40 Antenna

Pertinent design specifications for the AN/FRC-40 antenna are presented. Figure 14 defines the parameters for the Cassegrain system.

(1) Paraboloid

- D - 60 feet
- F - 25 feet
- F/D - 0.417
- Equation of parabola - $Y^2_\rho = 4\,FX = 100X$ (1)

(2) Hyperboloid

- d = 5.8 feet
- Equation of hyperbola $\frac{X + 7.46^2}{7.46} - \frac{Y_h^2}{6.29} = 1$ (2)

(3) Paraboloid Half Angle is given by

$$\tan \frac{\phi}{2} = \frac{D}{4F} \tag{3}$$

therefore $\phi = 62^o$.

(4) Feed Half Angle is given by

$$\tan \frac{a}{2} = \frac{D}{4MF} \tag{4}$$

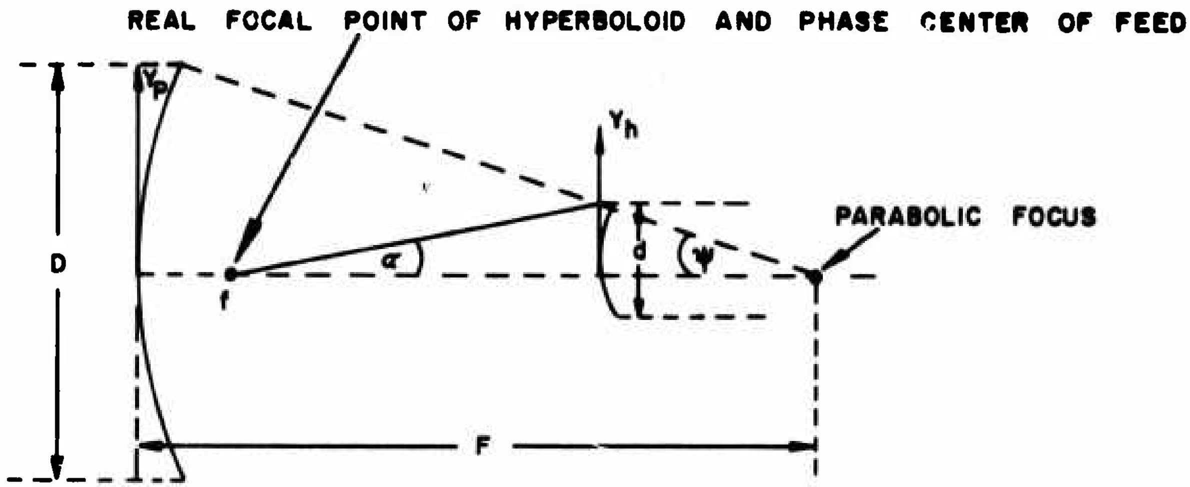


Figure 14. Parameters of Cassegrain Antenna

where M is magnification of Cassegrain system which is fixed by the existing design at M = 7.5, therefore $\alpha = 9.16^{o}$.

c. Estimate of Antenna Performance

The three parameters required to describe antenna performance are gain, sidelobe level and beamwidth. The specific design outlined above is herein analyzed to provide estimates of these parameters for the operating frequency of 3350 MHz + 250 MHz.

(1) Gain

The gain of an antenna is given by

$$G = \frac{4\pi A}{\lambda^2} \eta_t \qquad (5)$$

where A is the aperture area, λ is the operating wavelength and η_t is the overall antenna efficiency. The primary factors contributing to η_t in a practical design are:

η_a - Aperture efficiency

η_b - Blockage efficiency

η_{tol} - Surface tolerance efficiency

η_s - Spillover efficiency

Aperture efficiency - η_a - This factor expresses the reduction in gain below a uniformly illuminated aperture caused by a tapered illumination.

A typical aperture illumination for a parabolic reflector can be closely approximated by a circularly symmetric tapered distribution of the form[7]

$$f(r) = A + B\,(1 - \left(\frac{r}{R}\right)^{2P} \qquad (6)$$

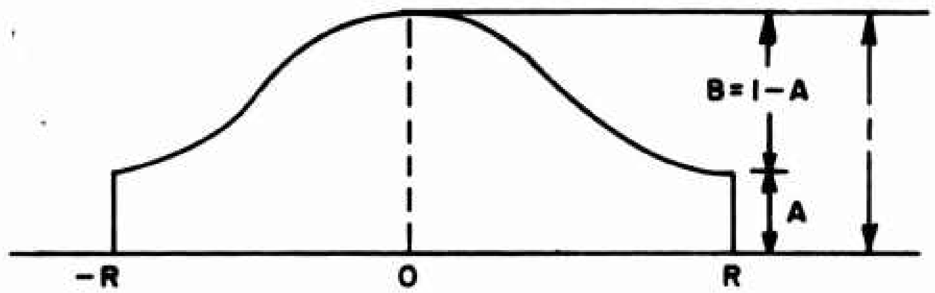


Figure 15. Aperture Illumination Function

where A + B = 1, and R is the radius of the aperture. This is shown in Figure 15. Assuming a -10 db edge taper (A = 0.3) and a value of P = 1.5, to provide a -24.8 db radiation pattern maximum sidelobe level, an aperture efficiency of 88.4%[8] is obtained.

Spillover efficiency - η_s - This factor represents the effectiveness of the feed to illuminate only the reflectors and minimize any energy which radiates elsewhere. Actual feed sources distribute energy in a pattern which tapers from a central maximum to nulls. If all this energy is intercepted by the reflectors, spillover loss would be zero; however, illumination efficiency would be rather low. As an example, a 3-db taper at the edge would achieve an η_s of approximately 50%[8]. An optimum value of taper which is common for the Cassegrain system is about -10 db illumination at the paraboloid edge. This value of taper yields a value of 85% for η_s. This value was obtained by integrating the total power contained in the universal feed pattern[9] and comparing it to the power contained within the -10 db points of this pattern.

Blockage efficiency - η_b - This represents the loss in gain caused by the presence of physical obstructions which shadow some portion of the radiated energy. In a Cassegrain system the contributors to blockage are the subreflector and its structural supports and the primary feed. The projection of the physical areas occupied by these structures on the aperture plane are the blockage factors (see Figure 16).

The calculations for blockage due to the subreflector and feed are straightforward and involve a ratio of their areas to the area of the paraboloid.

$$\text{subreflector blockage} = d^2/D^2 = 0.0094 \text{ or } 0.94\% \qquad (7)$$

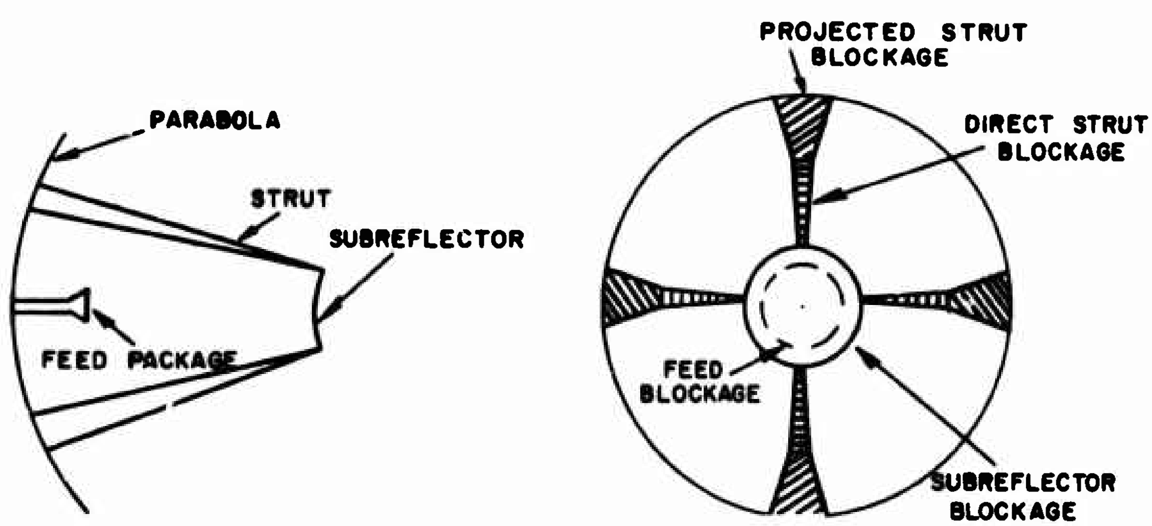


Figure 16. Aperture Blockage

The feed blockage should be considerably less than this and should fall within the shadow of the subreflector projection (see Figure 16). A combined value for the feed/subreflector combination is estimated as 1.5% of the total aperture area.

The blockage associated with the support struts consists of direct blockage and strut projected blockage (see Figure 16). In the existing design, the four struts intercept the reflector surface on a 24-foot diameter circle. The direct projection of these struts which taper from 1.5′ diameter at that point to 9″ diameter where they are attached to the subreflector results in a blockage of about 40 square feet on the aperture. The blockage due to viewing the struts from the paraboloid focal point (i.e., the four trapezoidal sections in Figure 16) project a total area of about 290 square feet on the paraboloid surface. This results in a total strut blockage of about 12 percent.

Since the various blockages shadow different portions of the parabolic dish, some of which are of relatively low illumination intensity (due to the assumed aperture illumination taper), the physical blockages calculated must be weighted to include illumination taper effects. Weighting both types of strut blockage the same and in the ratio of 1 to 2 with respect to the subreflector blockage, which shadows an area of peak intensity, results in a total effective blockage of 6.8% or $\eta_b = 93.2\%$.

Surface Tolerance efficiency - η_{tol} - This represents the loss in gain due to the RMS surface errors of the aperture and the resulting non-perfect focusing. The RMS surface tolerance of the parabolic primary reflector and the hyperbolic subreflector on the AN/FRC-40 antenna is reported as 0.040 inches and 0.030 inches, respectively. This results in a total RMS surface tolerance of 0.050 inches which at the high end of the operating frequency band (3600 MHz), representing the worst case, is 0.015 wavelength.

The degradation factor caused by antenna surface tolerance is given approximately by John Ruze[10], as

$$\eta_{tol} = \ell^{-16\pi^2 \left(\frac{\bar{\epsilon}}{\lambda}\right)^2} \tag{8}$$

where

$\bar{\epsilon}$ = RMS error = 0.05 inch

λ = free space wavelength

Letting $\lambda = 3.3''$ representing the high end of the frequency band and the worst case for surface tolerance analysis, $\eta_{tol} = 0.96$ or 96%.

Overall aperture efficiency is defined as the product of the contributing efficiency factors.

$$\eta_t = \eta_a \times \eta_s \times \eta_b \times \eta_{tol} \tag{9}$$

or

$$\eta_t = 65.6\%$$

Since we have not considered any detrimental effects of cross polarized loss, illumination phase deviation loss and subreflector diffraction loss, η_t will actually be lower than the above value. Estimating the combined effects of these losses as an 85 percent efficiency factor, overall efficiency can conservatively be estimated as $\eta_t = 55\%$. Using this value of η_t, equation (5) can be solved to determine the gain of the antenna.

$$G = \pi^2 \frac{D^2}{\lambda^2} \eta_t \tag{10}$$

Therefore

$$G = 194{,}500$$

or

$$G(db) = 52.8$$

where λ was taken as the wavelength at 3100 MHz, the low end of the operating band. Solving equation (1) for the high end of the frequency band yields a gain in db of

$$G = 54.2$$

These computations have been made to slide rule accuracy only.

(2) Sidelobe Level

In any design, the sidelobe levels are dependent upon the aperture illumination taper and the amount of aperture blockage present. The effect of increasing the taper across an aperture is to reduce the level of close in sidelobes and in general all sidelobes will be lowered. The effect of aperture blockage is to increase rather substantially the close in lobes.

A treatment of blockage effects on sidelobe levels presented by Jasik[11], and further developed by Jansen[12], yields the following expression which gives an approximation of how blockage affects sidelobes[12].

$$\text{New sidelobes} = 20 \log \frac{\frac{E_s}{E_m} - 2B^2}{1 - 2B^2} \qquad (11)$$

where E_s/E_m is the sidelobe-to-main beam voltage ratio for an unperturbed antenna and $2B^2$ is the blockage ratio for a tapered amplitude illumination. In Jansen's treatment[12], only the subreflector blockage was considered; however, in our case, the $2B^2$ factor will be more complex since it must also contain the effective strut blockage which must be further weighted to reflect the illumination taper. Estimates of the weighting are 1.8 for the direct strut blockage and 1.0 for the projected. Thus, equation (11) for an unperturbed sidelobe level of -24.8 db yields

$$\text{New Sidelobe} = 20 \log \left(\frac{0.056 + 0.085}{1 - 0.085}\right) \qquad (12)$$

$$= 20\,(0.149 - 0.966)$$

$$= -16.3 \text{ db}$$

Although this level of sidelobe radiation is not prohibitive for this operation, a lower sidelobe level would be preferable; nevertheless, it does not justify the costs involved in redesigning the strut supports which are the main cause of sidelobe degradation to enhance sidelobe performance. Furthermore, it is felt that this level could be decreased somewhat by using a more tapered aperture illumination in the feed design without degrading gain performance significantly.

(3) Beamwidth

The half-power beamwidth of an antenna is related to the size of the aperture and to the aperture illumination taper by the following relationship:

$$\text{Beamwidth} = K \frac{\lambda}{D} \tag{13}$$

where K is an aperture illumination taper constant. For the assumed aperture illumination

$$f(r) = 0.3 + 0.7 \left[1 - (r/R)^2\right]^{1.5} \tag{14}$$

The nominal value of K is 0.65(7). Thus, the existing AN/FRC-40 antenna is capable of providing a beamwidth on the order of 0.32° at the center frequency of 3350 MHz.

d. Summary

The preceding analysis provided the basis for predicting the antenna performance specifications wherein values for the various antenna parameters were obtained by a CW analysis. Although an analysis of the effects of broad bandwidth signals on antenna performance was not made, the CW analysis has been used to examine antenna pattern behavior as a function of frequency. It should be noted that the Cassegrain geometry provides equal pathlengths between the wavelength feed and the main reflector aperture, thereby preventing the transient "build up" of aperture illumination which destroys the broad band performance of many types of phased arrays.[13]

In terms of the gain at the peak of the beam, it is found that there is a gain variation of approximately 1.3 db across the band. Off-axis gain is further affected by frequency dependent beamwidth variation of 0.034° across the band as shown in Figure 17. Compensation for gain variation with frequency is possible in the system, but if this is done for on-axis gain, the gain variations at beam edges suffer additional deterioration. Results of gain variation are signal amplitude modulation which is not caused by target characteristics and are therefore undesirable. Since it is felt that the antenna will track well within the half power points of the main beam where a system compensated gain difference of less than 1 db is probable, the problem does not seem serious.

Sidelobe levels for a broad band signal will not be greater than the CW values. Moreover, the effective sidelobes should be lower, since at different frequencies the sidelobes are not geometrically coincident (Figure 17) effectively filling nulls and reducing peaks.

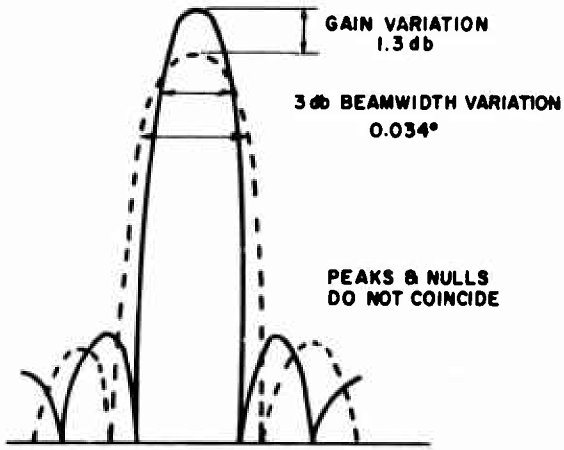


Figure 17. Pattern Variation with Frequency

Choice of antenna geometry can only partially solve the broad banding problem. Reflections from auxiliary dish and struts and coupled signals reflecting back from the monopulse feeds all occur within the 40-microsecond frequency swept pulse employed in the system and will affect the resulting waveforms. Off-axis waveform transmission even within the main beam has an additional source of deterioration due to the difference in path lengths to the target from different parts of the antenna aperture. Consider, for example, the signal from the edge of the dish compared with that from the center at an angle equal to the half power point of the main beam (one-sixth degree approximately). These signals travel path lengths to the target differing by one inch in length. Electromagnetic radiation in space travels at about 10^9 ft/sec or one foot per nanosecond, so the 40 microsecond pulse is 40,000 feet long. During this pulse the frequency sweeps 500 megacycles or 12,500 cycles per nanosecond. At this rate, the signal from the edge of the reflector differs in frequency by 1000 Hertz cycles compared with that from the center of the reflector. This effect, distributed over the entire aperture, must have its effects on the signal waveform. A study is being undertaken to further examine this problem.

2. Feed Design Consideration

The proposed radar must have an automatic tracking capability. This places certain requirements on the antenna and, more specifically, on the feed design. A preliminary evaluation of the tracking techniques available for a reflector antenna indicated that for the operating environment, a monopulse tracker would be most desirable. Moreover, a monopulse feed could be incorporated into the existing antenna structure with little, if any, modification to the existing antenna.

a. Comparison of Four-Horn vs Five-Horn Monopulse Design

In order to obtain azimuth and elevation tracking information, a minimum of four horns is required. In the transmit mode, the sum of the four horns provides the primary illumination. On receive, the difference patterns are used to derive the azimuth and elevation error signals for tracking while the sum pattern is used to derive a tracking reference signal plus the required target information (range, target signature, etc.).

A five-horn design is depicted in Figure 18. The center horn is used on transmit while all five horns are used on receive, but only the four exterior horns are used to derive the principal plane error signals. The center horn provides the tracking reference signal plus the target information.

The main advantage of the five-horn over the four-horn design is that data gathering and tracking are accomplished independently. Furthermore, the five-horn design offers a more simple design for the polarization requirements since the center horn is the only horn which requires orthogonally polarized outputs while tracking can be accomplished with singly polarized horns.

b. System Considerations Influencing the Five Horn Feed Design

The system requires that the feed be capable of operating under the following conditions:

- Power - 10 megawatts peak

  20 kilowatts average

- Frequency - 3350 MHz center frequency $\pm$ 250 MHz
- Polarization - Transmit linear vertical

  Receive linear vertical and horizontal simultaneously

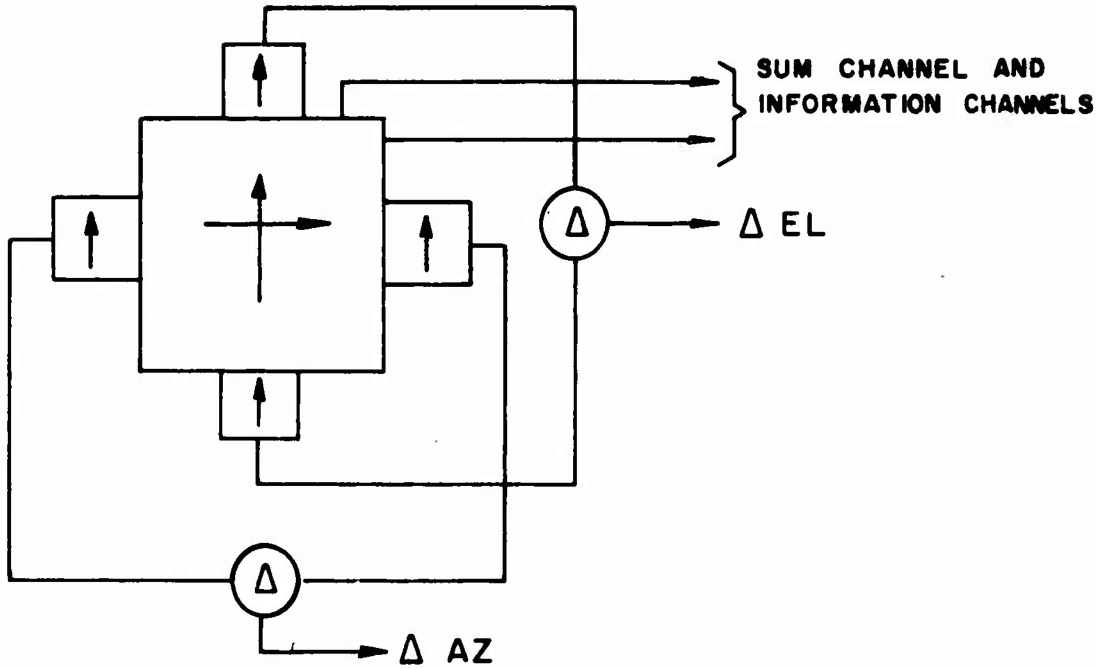


Figure 18. Five-Horn Feed Schematic

(1) Information Channels

The above system requirements pertain primarily to the transmit/data gathering functions and, because of the five-horn design, concern only the center horn. Since the proposed experiments are to obtain high resolution amplitude and phase profiles of space targets, extreme requirements have been imposed on data gathering channels. Data gathering is performed in the center horn which provides two orthogonally polarized outputs which form two independent information channels with the following characteristics:

(a) Amplitude Balance Between the Two Channels

Factors affecting the balance between channels other than mechanical tolerances are cross coupling between channels and mis-matches within channels. A 1.1:1 VSWR mismatch in one of the channels results in a 0.01 db loss in power. If cross coupling between channels or channel isolation is maintained at -40 db, an additional 0.08 db effect on unbalance is present. These conditions result in an unbalance which could be as low as 0.09 db. An assumed 30-db isolation between channels results in a total unbalance of 0.26 db. It is felt that the interchannel isolation can approach the design goal of -40 db and consequently a $\pm$ 0.1 db amplitude unbalance can be obtained.

(b) Phase Balance Between Channels

Phase dispersion differential existing between the two channels will consist primarily of dispersion due to mismatches and waveguide tolerances.

Phase dispersion caused by two mismatches in either channel can be represented by*

$$\phi = \tan^{-1} \frac{\sin 2\beta\ell}{\frac{1}{\Gamma_1\Gamma_2} + \cos 2\beta\ell} \qquad (15)$$

*See Appendix C

where

$$\beta = \frac{2\pi}{\lambda g}$$

and

λg = guide wavelength

and Γ_1, Γ_2 are the values of reflection coefficients of the two mismatches separated by a distance ℓ.

Since Γ_1 and Γ_2 are assumed small, then

$$\phi \simeq \Gamma_1 \Gamma_2 \sin 2\beta \ell \tag{16}$$

or

$$\phi \max \simeq \Gamma_1 \Gamma_2 \text{ (radians)} \tag{17}$$

As an example of differentiai phase dispersion consider two mismatches of 1.1:1 and 1.2:1, separated by 20 feet at the operating frequency, then ϕ max = 0.0045 radians or 0.27°. Since the overall specification on the feed horn is an input-to-output VSWR of 1.1:1, the effect of mismatches of this value will be less than that given in the example. Furthermore, the effect of more discontinuities of smaller magnitude spread along the waveguide channels will result in an even lower phase dispersion ripple.

If waveguide tolerance in the broad dimension of the guide can be kept to a differential of 0.010 inches, a phase differential dispersion of 0.8° will result. The sum of the two dispersions obtainable in this example by combining waveguide tolerance and mismatch is 1.07° differential dispersion. An interchannel phase balance of ±3.0° should be met without difficulty in this case.

(c) Insertion Loss

The overall length of the feed system (i.e., from mouth of the horn to the transmit receive enclosure) will be approximately 35 feet. Since maximum attenuation in RG/48U waveguide is less than 1 db/100 feet across the operating band, insertion loss in each channel should be about 0.3 db.

(d) Pressurization

The center horn or at least the transmit channel of the feed must be pressurized to prevent breakdown of the 10 MW peak power. For RG/48U waveguide, the required pressurization is about 15 psig.

(2) Tracking Channels

The tracking horns can be designed independently to operate in the vertically polarized mode only over a 150-MHz bandwidth. The performance of the tracking horns can best be discussed in terms of the monopulse difference patterns.

The difference patterns are generated by taking the difference of the two symmetrically squinted beams (see Figure 19) generated by the appropriate pairs of tracking horns. Also shown on the figure is the sum pattern of the squinted beams.

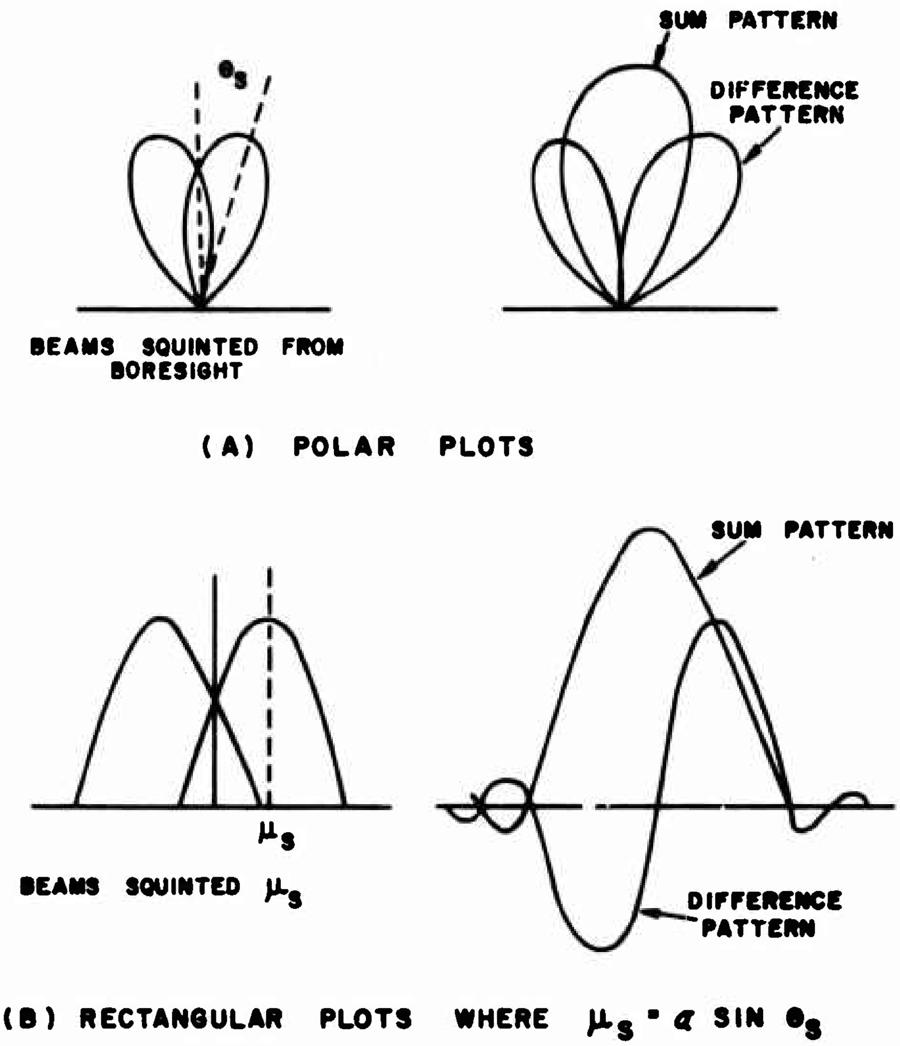


Figure 19. Monopulse Sum and Difference Patterns

The characteristics of the difference pattern necessary to describe performance are gain, sidelobe level, null depth and difference pattern slope.

(a) The gain and sidelobe level of the monopulse pattern are interrelated and highly dependent on the particular antenna design. If one were to optimize the gain of the difference pattern alone, he would do so at the expense of high sidelobes. Since the AN/FRC-40 antenna has excessive aperture blockage, sidelobes would be excessively high in the optimum difference pattern gain design, consequently a gain sidelobe level trade off must be considered.

Assuming a 1 to 3 db cross over of the squinted beams, the gain of the difference pattern peak could be as high as 2.47 db[14] below the sum pattern peak in the optimum gain design. The illumination required to provide this gain would result in excessive spillover and unperturbed sidelobes of the order of -18 db. It was shown previously

that the aperture blockage present results in about an 8-db degradation of the sidelobe levels in the sum pattern. Consequently, optimum gain design would have prohibitive sidelobe characteristics. An aperture distribution or primary illumination which reduces spillover would result in a lower sidelobe level. It is estimated that the antenna design with proper illumination could provide final sidelobes of 17-18 db below the peak of the difference pattern. This would be accompanied by a loss in gain but this should be kept within -6 db of the peak of the sum pattern.

(b) Null Depth, Width and Error Signal Slope

The error signal slope may be determined from the relationship given by Kinsey[14] for a circular aperture,

$$K_o = \frac{\sqrt{G_o}}{2}\ \text{Volts}/_{\text{rad}} = \frac{\sqrt{G_o}}{114.6}\ \text{Volts}/_{\text{degree}} \tag{18}$$

where

K_o = Maximum error signal slope

G_o = Maximum theoretical sum mode gain

$$\frac{4\pi A}{\lambda^2}$$

$$K_o = \frac{\frac{\pi D}{\lambda}}{114.6} = 5.6\ \frac{\text{difference channel volts}}{\text{degree target offset}} \tag{19}$$

This value must be corrected to account for the actual sum mode gain (i.e., consider 55% efficiency) and the deviation from maximum slope resulting from the actual difference aperture illumination function which results in a factor of about 0.8[15],

or

$$K \simeq (5.6)\left(\sqrt{0.55}\right)(0.80) = 3.4\ \text{volt/degree} \tag{20}$$

(3) The width of the null at a specified level below the sum channel peak can be determined from the slope of the error signal. Assuming a null depth of -35 db (0.0178 volts in difference channel relative to one volt in the sum channel), the full width of the difference null will be approximately

$$\theta_n = \frac{2(0.0178)}{3.4}\ \frac{\text{volts}}{\text{volts}/_{\text{degree}}} = 0.0105^{\circ} \tag{21}$$

B. TRANSMITTER DESIGN CRITERIA

The primary design criteria for the transmitter are based on the final performance expected and the economics of using the presently available facilities.

In order to maintain very low time sidelobes on the compressed pulse, consideration must be given to the phase-frequency characteristics and amplitude-frequency response of the transmitter. The ideal phase performance of the transmitter chain would be a linear phase-frequency characteristic with no high frequency components of phase

deviations. In reality, this is difficult to achieve over the very wide bandwidth of the system. The number of individual devices used in the RF chain also contributes to a high frequency perturbation because of interstage coupling, reflections due to mismatch, and nonlinearities of the individual devices. The operating conditions and environment of each stage will add another source of phase nonlinearity. Coaxial cables, connectors, waveguide sections and filters will produce a phase response dependent on the relative lengths involved, mismatch and loss as a function of frequency. Each of the components will have to be designed and error values assigned to each in order that the system as a whole will meet the final parameters.

The transmitter chain will consist of a minimum of two traveling wave tube amplifiers in cascade driving a hybrid traveling wave tube - klystron (Twystron®) final power amplifier. The initial or pre-driver will serve as a phase compensating element for the entire chain as well as amplify the input signal. The second TWT must provide sufficient output power to drive the hybrid TWT and overcome the insertion losses in an isolator or attenuator located between the driver and the final amplifier.

Each of the three tubes must be operated in its minimum phase distortion mode. The phase response of each individual tube is dependent on certain intrinsic properties of the tube itself and also on external factors which cause a reaction on the tube.

Other than designing and building the tube to have the desired phase characteristics, nothing can be done to correct the tube phase performance. However, those external variables that influence the phase-frequency response of the tube can be controlled. These are the external factors which tend to change the electrical length of the tubes. Beam voltage pulse variations which cause stream velocity modulations produce a phase modulation. Electron current density variations cause amplitude modulations in the output which are accompanied by phase modulation components. Magnetic field beam focusing variations, filament voltage variations, magnetic effects of the filament voltage on the beam current, drive level variations, output VSWR, and a host of other parameters all contribute to the combined effect on the tube behavior.

To achieve the desired degree of phase linearity, each of these variables will have to be closely controlled.

The modulator required to drive the final power amplifier will play an important part in determining the phase-frequency response of the hybrid TWT. In addition to providing an adequate peak and average power level to operate the final amplifier, the pulse shape must be carefully controlled in order to enhance the phase-frequency response of the tube.

The ideal pulse shape would be of constant amplitude during the time that the RF drive is applied to the input of the final amplifier. Any droop or ripple voltage on the pulse will cause a phase perturbation.

The type modulator that can be used to drive the final power amplifier is restricted by the mechanical and electrical parameters of the existing equipment. The location of the high voltage power supply in the operations building requires a high voltage cable to conduct the high voltage to the antenna enclosure. Because of its size and weight, locating the power supply on the antenna itself would be impractical.

Another factor affecting the type of modulator is the type of beam control necessary for the hybrid TWT. A series beam switch using a hard tube modulator could be used except for a very important factor. The pulse current from the power supply to the load would have to go through the high voltage cable which is approximately 250 feet in length. The inductance and capacitance would be shock excited, causing a severe RFI problem to other equipment. In addition to the long length of cable involved, the choice of high voltage switch tube to use in the modulator would be another risk. A large high voltage switch tube is nearly always designed to be operated with the anode-cathode axis in a vertical plane. Because of the moving antenna platform, the filaments and other elements of the

---

Registered Varian Trade Mark.

tube would be subjected to a varying orientation which could cause interelectrode shorts in the tube.

Another approach for the modulator design would be to use a modulating anode for beam control of the hybrid TWT and pulse the anode with a floating deck hard tube modulator. Here again, the power supply is the limiting factor. The existing power supply has insufficient output voltage to operate the hybrid TWT. There would be a greater problem with the space available in the antenna enclosure to handle the higher d-c voltages associated with a modulating anode type of final power amplifier tube and modulator.

Another problem associated with the two preceding modulators would be the high voltage switch tubes and protection of the load and modulator in case of arcs in the system.

A line type modulator to pulse the hybrid TWT is another possibility. It can operate from the existing power supply. The peak current on the high voltage cable is low and also has a rather long rise and fall time (greater than seven milliseconds). The low peak currents and low rise and fall times would reduce the RFI problems associated with the long high voltage cable. The components associated with a typical line modulator are relatively compact and sturdy. The maximum high voltage in the line modulator would be less than 40 kv except at the cathode of the hybrid TWT and then it is contained in an enclosed pulse transformer tank. The energy that can be dumped into a load that is arcing is limited to that contained in one pulse, provided that the trigger that would initiate the next pulse is inhibited. This is so because the PFN and pulse transformer serve as an isolator between the high voltage power supply and the final load. The main limiting factors of the line modulator are the impedance requirements, variable pulse lengths, and pulse top ripple.

The line modulator must be able to accommodate two different final power amplifier tubes (Phase I and Phase II). This imposes a requirement for dual pulse lengths, different peak and average power operation and slightly different impedances in the two conditions. It appears that using dual pulse forming networks is the most reliable method to achieve all of the required conditions.

The required cathode pulse flatness is a rather difficult problem. In order to reduce the phase modulation as a result of cathode voltage ripple, 0.03% droop and ripple on the voltage pulse are desirable. Producing a pulse with this flatness is difficult, but measuring the pulse can be just as difficult. It is unlikely that the pulse flatness can be achieved by passive components alone. An active pulse level control is required that has sufficient control bandwidth to correct for high frequency ripple components. Such a device could be a variable load which varies its impedance in order that the line modulator looks into an exactly matched load condition during the pulse. The variable load would be in series or shunt with the fixed tube load and vary so as to remove the ripple voltage during the pulse.

Achieving a linear phase frequency transfer characteristic from the entire transmitter is also a difficult problem. After all the external factors which can cause the phase-frequency nonlinearities are minimized, the tube phase parameters and the inherent dispersive components in waveguide runs still remain to be concerned with. A phase compensating circuit would thus be desirable. It is unlikely that a passive component could be versatile enough to compensate for high frequency components of phase ripple, but could help to remove a low frequency component. Low frequency in this case would be defined as one or two cycles of phase deviation across the entire bandwidth. For the phase perturbations which are stable and are of a high frequency nature, an active compensating circuit is necessary. The bandwidth and the gain of the compensating circuitry must be able to correct for the highest frequency component of interest across the bandwidth.

The phase response of the transmitter on a pulse-to-pulse and on an intra-pulse basis must first be determined in order that compensating techniques can be applied. Several measurement techniques are potentially able to fulfill the requirements of the system. The measurement accuracy must be at least as accurate as the required linearity requirements.

The maximum allowed phase nonlinearity of the transmitter would thus require an accuracy in measurement of at least ± 10° at any point in the pulse width. For more complete analysis, an accuracy of less than ± one degree is desirable. This accuracy figure must include all of the cables, connectors, and components. In lieu of a direct phase measurement, an alternate approach can be taken by adjusting and correcting the phase and amplitude performance of the transmitter until the desired time sidelobes are obtained. This is, in reality, the ultimate objective of phase correction. This approach will be investigated by using a manual form of compensation by adjusting the phase of the transmitter at one half microsecond intervals across the pulse. Thus, for each six-megahertz interval of bandwidth, an average phase correction will be made by an adjustable input voltage.

C. HIGH VOLTAGE POWER SUPPLY AND COOLING SYSTEM

The existing high voltage supply and cooling system was manufactured by Energy Systems, Inc. The high voltage power supply is located in a portion of the operations building (Figure 6) and the cooling system is located in a separate building adjacent to the radar pedestal.

The system consists of a high voltage power supply capable of one megawatt d-c at voltages between 20,000 and 120,000; a cooling system capable of removing up to 1.2 megawatts of heat from the electronic equipment and transferring it to the outside air; a d-c dummy load capable of fully loading the high voltage power supply; a crowbar system to provide fault diversion for fast load protection; a control console, a-c primary line power control and monitoring; and low level circuits, controls and power supplies.

The high voltage supply consists of a 4160-volt circuit breaker, a full range Inductrol, vacuum switches for primary power control, a three-phase beam transformer with two separate sets of secondaries, two complete three-phase full-wave silicon rectifier bridges, and a capacitor bank. Continuous adjustment of the high voltage output is provided by the Inductrol. Rated power over the specified voltage range is obtained by reconnecting the power supply in four configurations. These configurations include connecting the transformer secondaries in wye or in delta and connecting the d-c outputs of the two sets of rectifiers in series or in parallel.

The crowbar system contains a triggered arc gap, a triggering and logic mechanism, and a surge resistor which absorbs the energy stored in the capacitor bank when the crowbar is activated.

The capacitor bank consists of six groups of capacitors arranged in three levels. The bottom two groups and the top two groups are always tied in parallel. The middle two groups are floating. The different capacitor bank arrangements are obtained by different interconnections between the floating level two and the other levels by means of various capacitor links.

The high voltage room is a non-livable area whose entrances are protected by key interlocks.

The specifications for the high voltage power supply are:

- Input Power: 4160 volts ± 5%, 60 Hz ± 5%, 3 phase, 210 amperes and 1400 Kva by rated output, 5% maximum unbalance between any two phases.
- Output Power: 1 megawatt d-c between 20,000 and 120,000 volts, obtainable in four voltage ranges by alternate transformer and rectifier connections.
- Delta-parallel, 0-35 kv. Current rating 50 amperes at 20 kv to 28.9 amperes at 35 kv.
- Wye-parallel, 0-60 kv. Current rating 28.9 amperes at 35 kv to 16.7 amperes at 60 kv.
- Delta-series, 0-70 kv. Current rating 25 amperes at 40 kv to 14.4 amperes at 70 kv.
- Wye-series, 0-120 kv. Current rating 14.4 amperes at 70 kv to 8.3 amperes at 120 kv.

- Polarity: Negative or positive.
- Ripple: Less than 0.5% peak-to-peak with the full capacitor bank for each connection and up to full load current.
- Power Supply Impedance: 18% reactance including source (based on source impedance 0.0297 +j 0.0898 on 3 MVA base).
- Output Voltage Adjustment: With motor driven Inductrol, 4160 volts input ± 5%, 7500 volts output, 90 second motor runup time, control power 120 volts, 60 Hz, single-phase.
- Inductrol Regulator: Regulates output voltage to ± 1% over the range of 1875 to 7500 volts output (25% to 100% of output).
- Capacitor Bank: 500,000 joules energy storage capability, available in three voltage ratings.
  - 625 microfarads at 40 kv
  - 156 microfarads at 80 kv
  - 69.5 microfarads at 120 kv

The cooling system consists of a distilled water loop, a brine loop and a glycol loop. The distilled water is used to cool the electronic equipment. This medium is used in order to provide the heat capacity and maintain the high rate of heat transfer required in the Twystron collector. The glycol loop is used to transfer the heat from the distilled water to the ultimate sink, the outside air. The glycol loop provides isolation between the air and the distilled water to reduce the danger of freezing the water. The brine loop is used to cool the d-c dummy load, as well as to furnish the resistive element for the load. The heat in the brine is transferred to the distilled water, then to the glycol, and then to the air.

Separate distilled water pumps are used to cool the transmitter and the brine loop. The pump for cooling the brine loop has sufficient flow rate for the one-megawatt capability of the d-c dummy load.

1. D-C Dummy Load Loop

- Absorb and remove one megawatt of d-c power from d-c dummy load resistor.
- Flow rate: 300 gpm.
- Maximum temperature leaving dummy load: 82°C.
- Surge tank capacity: 250 gallons (200 gallon liquid).
- Salinity: can be increased in steps from the main control console.
- Temperature control of input brine to DCDL by thermostatically controlled bypass valve around the brine heat exchanger.

2. Distilled Water Loop

- Transfer up to one megawatt of heat from brine loop to glycol loop.
- Transfer up to 400 kw of heat from transmitter to glycol loop.
- Rated to transfer up to 1.2 megawatts of heat from transmitter to glycol loop by changing to a 400 gpm transmitter cooling pump.

- Flow rate: 416 gpm when cooling brine; 160 gpm at 120 psi when cooling transmitter during SYSTEM ON; 45 gpm at 14 psi to transmitter during SYSTEM OFF.
- Temperature: Temperature controlled at inlet to surge tank by a pneumatic control system to maintain system temperature of 132 ± 2.5°F during HIGH VOLTAGE ON: 120°F during SYSTEM ON-HIGH VOLTAGE OFF; 45°F during SYSTEM OFF.
- Uses hot water from building heating system to maintain cooling water temperature; 670,000 Btu/hr from 20 gpm of hot water at 180°F is required.
- Surge tank capacity: 400 gallons (250 gallon liquid).
- Storage tank capacity: 1000 gallons (800 gallon liquid).
- Purification loop to filter and repurify the distilled water used for cooling.
- Operation of slow speed transmitter pump controlled by outside ambient temperature. Pump operates below 35°F ambient temperature to circulate warm water through the system.
- Check valves used throughout system to eliminate need for valve adjustments when changing cooling between d-c dummy load and transmitter.

3. Glycol Loop

- Transfers up to 1.2 megawatts from distilled water loop to air at 100°F maximum ambient temperature.
- Flow rate: 416 or 832 gpm, selectable.
- Glycol temperature: 120°F ± 5°F, thermostatically controlled.
- Glycol mixture: 60 by volume of ethylene-glycol PM1717 in water which has been treated to remove hardness.
- Glycol surge tank capacity: 500 gallons.
- Radiators: Three radiators with 10-foot diameter fans; individually controlled by temperature switches to start at glycol temperatures of 119°F, 124°F, 129°F.
- Pressure relief valve around radiators prevents pressures in excess of 35 psi.

4. D-C Dummy Load

- Power rating: One megawatt d-c maximum.
- Voltage rating: 120 kv d-c maximum.
- Current rating: 50 amperes d-c maximum.
- Loss medium: Brine solution. Salinity can be increased in steps from control console.

Both the power supply and the heat exchange system have more than adequate capacity to meet the needs of the proposed transmitter design. The cooling system has a well stabilized temperature control characteristic which would prevent excessive phase-frequency drift characteristics as a function of temperature.

D. RECEIVER DESIGN CRITERIA

The requirements for the receiver subsystem, as specified under the original design goals, are based upon Phase II of the proposed program. This phase of the task required a 500-MHz instantaneous bandwidth centered at 3350 MHz having a maximum power handling capability of 10-megawatts peak, 20-kilowatts average, a maximum noise figure of 6.5 db and a total phase nonlinearity of ± 5°.

This project required limiting to the consideration of "off-the-shelf" microwave devices in order to avoid costly and time consuming developments. Technical discussions were conducted internally and with industry regarding present state-of-the-art devices. Based upon these discussions, it was considered that existing components are available which would achieve the desired design goals.

The receiver subsystem consists of four separate channels: two broadband information channels, called the "vertical" and "horizontal" channels, and two narrow band monopulse error channels. The vertical channel is the high power channel while the horizontal channel has 30 db of isolation from the vertical channel in the transmit state. Both the vertical and horizontal channels must be electrically identical.

A survey was conducted to determine the optimum parameters involved in obtaining a receiver subsystem utilizing present state-of-the-art microwave devices. The components to be investigated were low noise receivers, high power duplexers, high power and low power broadband couplers, low power narrow band couplers, wideband and narrowband gated limiters, mixers and IF strips.

The major areas that demanded prime consideration were the high power duplexer and the low noise receiver elements. A competititve analysis was not required of the other electronic devices as they were available components dependent only on electrical parameters that could be defined as applicable to the system's requirements.

Two types of high power duplexers were analyzed. The ferrite and gaseous type duplexers were considered the most optimum approaches for achieving the specified design goals of the work statement. A comparison of the merits of the solid state duplexer relative to the gaseous type duplexer resulted in the following determinations:

(a) The ferrite type duplexer is capable of handling the high power requirements of the specifications. Assuming no peak or average overload conditions, the duplexer would have an indefinite life time, no recovery time involved, no replacement time required and lower operating cost because of the long life time expectancy.

(b) Although the gaseous type duplexer had the high peak and average power capability, there appeared to be several disadvantages as compared to the ferrite duplexer. This device had a guaranteed life of only 1000 hours. The recovery time of the gaseous type duplexer depends upon the diffusion and recombination processes. This limits the recovery time of the device dependent upon the frequency, geometry and the plasma media. There is, therefore, a replacement time problem and increased operating cost considerations as opposed to the ferrite duplexer approach.

Based upon this preliminary data, it was considered that the optimum approach for satisfactorily complying with the specifications was to proceed in the direction of a ferrite type duplexer. However, subsequent evaluation of the operating requirement for electrically identical vertical and horizontal channels could not be achieved by a ferrite duplexer. The phase tracking of $\pm 1.0^{\circ}$ and the insertion loss tracking of 0.1 db required for the receiver subsystem were considered beyond the state of the art for ferrite duplexer devices. As these two parameters are essential for obtaining the necessary high resolution amplitudes and phase profiles dictated by system's requirements, the gaseous duplexer approach was given primary consideration.

Low noise receiver devices evaluated were vacuum tube RF amplifier, transistor amplifier, crystal mixer, MASER, traveling wave tube, parametric amplifier and tunnel diode amplifier. The critical technical areas that necessitated major consideration were receiver sensitivity, frequency range, bandwidth, gain, gain tracking, phase tracking, volume and weight requirements.

The vacuum tube and transistor amplifiers were limited by poor noise performance and therefore not considered compliant with the noise figure requirements. The MASER, although having ultra low noise properties, could not attain the bandwidth and dynamic range specifications. In the case of the parametric amplifier and traveling wave tube, the cost was a prohibitive factor.

The tunnel diode low noise receiver appeared to be the only electronic device that could fulfill the overall system's requirements within reasonable costs. The noise figure and bandwidth can be successfully achieved by incorporation of the low noise tunnel diode amplifier in the receiver subsystem channels. These devices are small in size and light in weight and can be supplied in weatherproof cases for antenna mounting, if necessary, as well as in dustproof or rack-mounted configurations. Since these units are operated with low d-c voltages, they have extremely stable gain and phase characteristics. The tunnel diode amplifier can be subjected to any arbitrary loading impedance at the input and output simultaneously, without causing any spurious oscillations either in or out of the frequency band of interest. The environmental characteristics of these devices are superior in comparison to other competitive low noise devices.

In summary, analysis indicated that the optimum approach for achieving the design parameters was the gaseous duplexer configuration and it was determined that the best approach for resolving the receiver design goals was the utilization of a low noise tunnel diode amplifier. The incorporation of these two microwave devices should be capable of achieving optimum performance of the receiver subsystem.

E. SIGNAL PROCESSOR

The development of the signal processor specifications has been accomplished with enough generality to permit a variety of techniques to be considered, yet specific enough to define the parameters of other system components. Because of the simultaneous requirement for high range resolution and high transmit power, it is immediately obvious that some form of pulse compression must be used. Transmitter characteristics and other system requirements, such as data rate, information bandwidth and range window length leads to a fixed transmit pulse length of 20 μsec for Phase I and 40 μsec for Phase II. The effective pulse compression ratios are the transmit pulse length times the information bandwidth, which for Phase I is 5,000 and for Phase II is 20,000.

The general requirement of the signal processor is to generate the basic wideband expanded waveform and to process the received signal by compressing and translating the information band so that the received signals can be recorded. The radar output parameters of interest will be the amplitude and relative phase response of a target (one-foot nominal resolution) within a 100-foot range window. The orthogonal polarization response is also obtained and processed through common channels. The basic form of the signal processor is depicted within the dotted lines of Figure 20. To accomplish the general requirements, the signal processor should function as follows:

The exciter must provide a suitable pulse expanded waveform to the transmitter for high power amplification. The waveform should include the necessary compensation factors to take into account the distortions imposed upon it by the effects of doppler, ionosphere and microwave components. To alleviate the microwave distortion problem the wide band portions must be located as close to the feed horn as physically possible.

Upon reception of the wide band signal, narrow banding techniques should be used immediately to ease signal handling difficulties and further reduce distortion which may otherwise occur. A linear FM pulse compression technique utilizing active local oscillator cross-correlation to accomplish this function is used. The local oscillator is synchronized and cross-correlated with the received expanded signal, effectively reducing the signal bandwidth. Range resolution is maintained as a function of the resultant correlation frequency. Such narrow banding also permits the cross-polarized signals to be relatively delayed and processed through a common channel, thus simplifying the subsequent signal processing.

Of utmost importance is the ability to slow down the signals, i.e., reduce the 250 MHz (or 500 MHz) information bandwidths to permit recording by known A/D converter techniques. It should be noted that only 100 feet (0.1 μsec) of the total range is to be recorded while the PRF interval (data rate) is 1/70 sec. Narrow banding of the signal by this method provides a time expansion characteristic, making it possible to use the long interval between pulses to record the signal at a much slower rate than is indicated by the overall 250 MHz (or 500 MHz) bandwidths

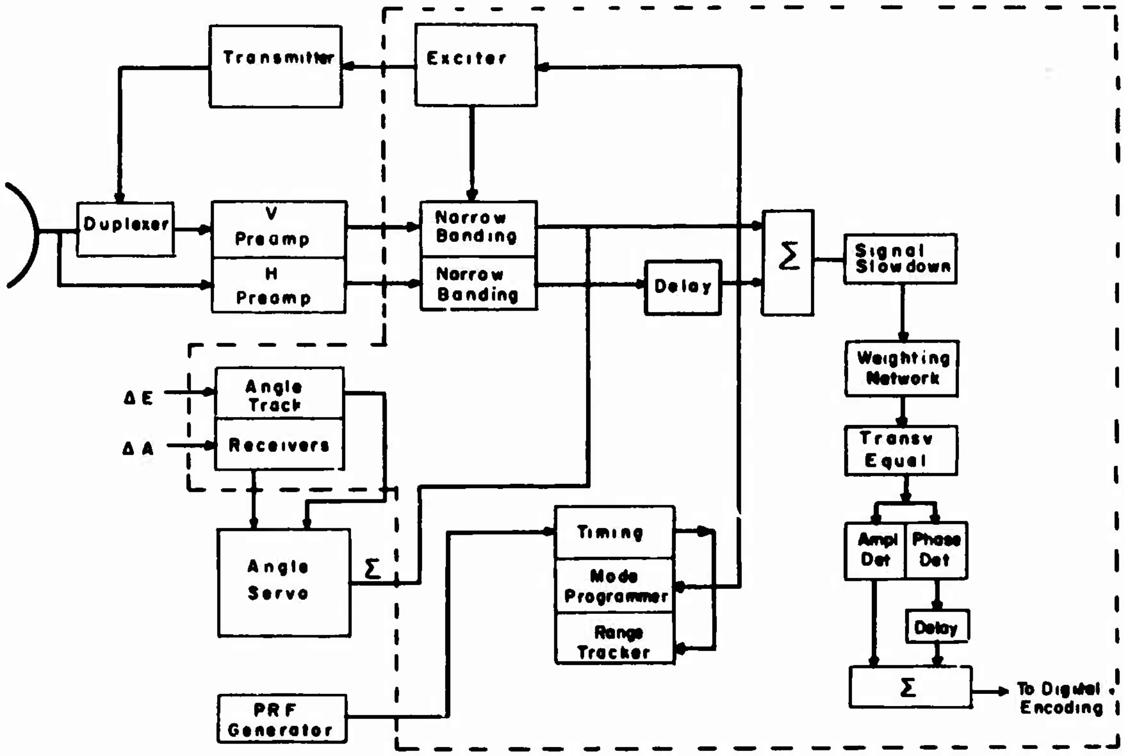


Figure 20. Simplified Basic Design

involved. The desired time expansion will be provided in part by the signal processing technique. It can be accomplished as shown in Figure 8, for example, by the use of dispersive delay lines. These lines when used in conjunction with the frequency modulated mixer can simultaneously compress the waveform and delay the signal in accordance with the differential range (frequency). High speed sampling, A/D conversion, and interim buffer storage with reduced rate readout provide additional time expansion for reliable tape recording.

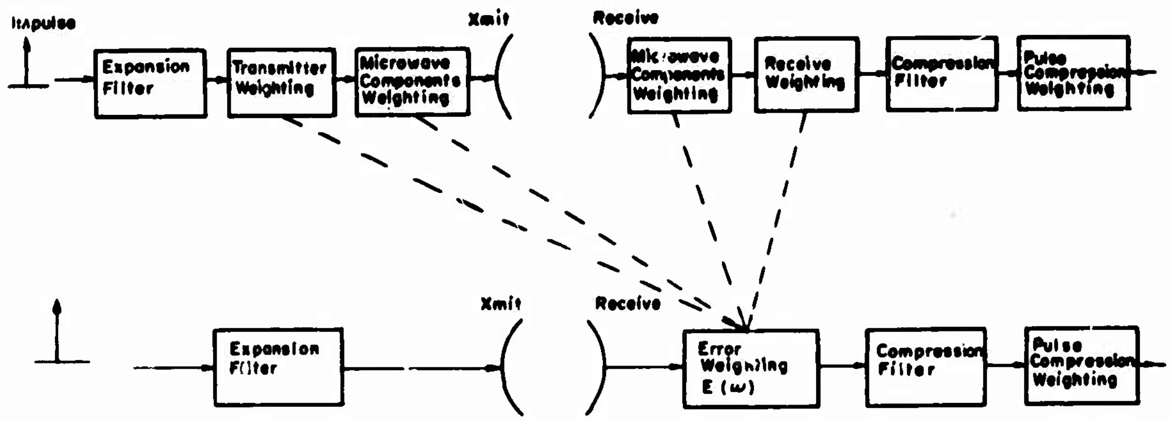
The weighting for pulse compression signals is also indicated in Figure 20; however, it must be realized that all components throughout the system need to be considered as contributing to the resultant weighting. This is indicated in Figure 21 and will be discussed later. In the same sense as signal weighting, a transversal equalizer should be incorporated to adjust the overall signal response after reception.

The final segments of the signal processor are the amplitude and phase detectors which will present the analog signals to the data handling equipment. Figure 20 indicates amplitude and phase outputs; however, it is realized that the quadrature components of the signal may also be used.

Figure 20 also indicates functions such as angle track receivers, timing, mode programmer, and range tracker, which are to be included in the overall signal processing package to effect a complete modification of the Floyd Site.

An overall system response to result in the signal range sidelobes of -35 db is considered feasible if due care is given to the system design, exclusive of atmospheric effects. The low sidelobe response requirement is a result of the primary function of the radar, i.e., satellite observation. Various investigations[16] on short pulse scattering of an object indicates that at least a 35-db dynamic range is desirable to distinguish small scatterers from sidelobes of large scatterers in the same range window.

A. Elementary Block Diagram Of Weighting Errors



B. Equivalent Block Diagram Of The Weighting

Figure 21. Amplitude & Phase Errors

In fact, the Fourier transform of the power spectrum is the time response of the pulse compression system. With a signal spectrum $S(\omega)$, and a matched filter response $H(\omega) = S^*(\omega)$, the output spectrum of the signal is

$$P(\omega) = S(\omega)\, H(\omega) = |S(\omega)|^2 \qquad (1)$$

which is also the power spectrum of the input signal. The Fourier transform of $P(\omega)$ is the output signal. It is immediately evident from Figure 21 that all system components will contribute to the resulting signal spectrum (amplitude and phase). The output can be represented by the Fourier transform of $P(\omega)\, E(\omega)$ where $E(\omega)$, for a linear network, represents the phase and amplitude effects.

In Figure 21, a completely linear system is assumed. However, the technique for narrow banding signals, as shown in Figure 3, is not necessarily linear. Small nonlinearities in the ramp can be shown to give phase errors equivalent to a linear network which may then be treated using paired echo theory. In addition, nonlinearities in the ramp may affect the magnitude of the error as a function of target position within the range window. This effect is discussed later. In general, a completely linear ramp will not produce errors; therefore, the linear system model can be used.

Assuming tolerance can be held to small values, the required tolerances can be estimated, using paired echo theory[17]. The power of this theory lies in its ability to separately express the effects of amplitude and phase distortion in terms of the original undistorted time function corresponding to the spectrum of interest (before distortion) and various time delayed replicas resulting from the distortion permitting one to determine the resultant overall distortion by direct summation of these similar temporal waveforms.

Since the paired echo analysis utilizes a Fourier Series to express the amplitude and/or phase distortion characteristics, the overall distortion due to even non-sinusoidal amplitude and/or phase variation can be analyzed in terms of a number of paired echoes of the original undistorted waveform, simplifying the problems of analysis tremendously.[18]

Making use of this theory, while assuming sinusoidal perturbations, it can be shown that the peak amplitude of the spectrum is increased by a factor of 1.04 corresponding tc a peak amplitude ripple of 0.34 db for -34 db time sidelobes. Similarly, the peak phase error ripple is determined to be $\pm$ 2.3 degrees.

These errors must be constrained throughout the entire system if a nominally -35 db sidelobe response is to be realized.

It is anticipated that the transmitter signal transfer characteristic will not in itself meet these requirements. In addition, it has been found difficult to obtain reliable phase data from tube manufacturers from which the transmitter contractor may design. For this reason it is believed desirable to have active correction loops in the transmitter and signal processor to aid in achieving these requirements. It is believed that a transversal equalizer will then effectively reduce the sidelobes to the desired level.

All wide band portions of the system should be located near the feed horn. The effects of waveguide dispersion and VSWR on the signal are discussed in Appendix C.

The slowed down output information IF bandwidth should be 2.5 MHz. This bandwidth is selected for several reasons. First, an output signal must be specified as a basis for A/D conversion. It also appears reasonable to use with delay lines. If dispersive delay lines are to be used in the slow down technique, the time bandwidth product must be considered. The 2.5 MHz bandwidth results in a 50 to 1 and 100 to 1 time bandwidth product for Phase I and Phase II, respectively. As discussed in the Data Handling Section, the 2.5 MHz bandwidth appears feasible to A/D convert and record.

The signal processor should operate in various modes for angle and range acquisition and tracking. These modes are described elsewhere.

A range window of 100 feet has been selected. This is a trade off between the expected target lengths, resolution, data handling capabilities and accuracies involved. From the point of view of minimizing the effects of nonlinearities in the L.O. ramp, as short a window length as possible is desired. For the purpose of illustrating this effect, we will represent the frequency vs time curve as the difference frequency resulting from the mixer operation on two signals as shown in Figure 22A. If there were no time shift in the range window due to target position, the ideal mixer output would be a constant difference frequency, as shown in Figure 22B. However, target signals from different portions of the range window have a relative time shift in the range window with respect to the L.O. signal as shown in Figure 22C. If the frequency function of the target signal vs time is represented by

$$f_r = f_o + k_1 t + k_2 \sin 2\pi f_m t \tag{2}$$

where

$$t_1 < t < t_2$$

and

$$f_{LO} = (f_o - f_{IF}) + k_1 (t - \Delta t) + k_2 \sin 2\pi f_m (t - \Delta t) \tag{3}$$

is the local oscillator signal with an arbitrary time shift. After mixing, the difference frequency can be represented by

$$f_r - f_{LO} = f_{IF} + k_1 \Delta t + k_2 [\sin 2\pi f_m t - \sin 2\pi f_m (t - \Delta t)] \tag{4}$$

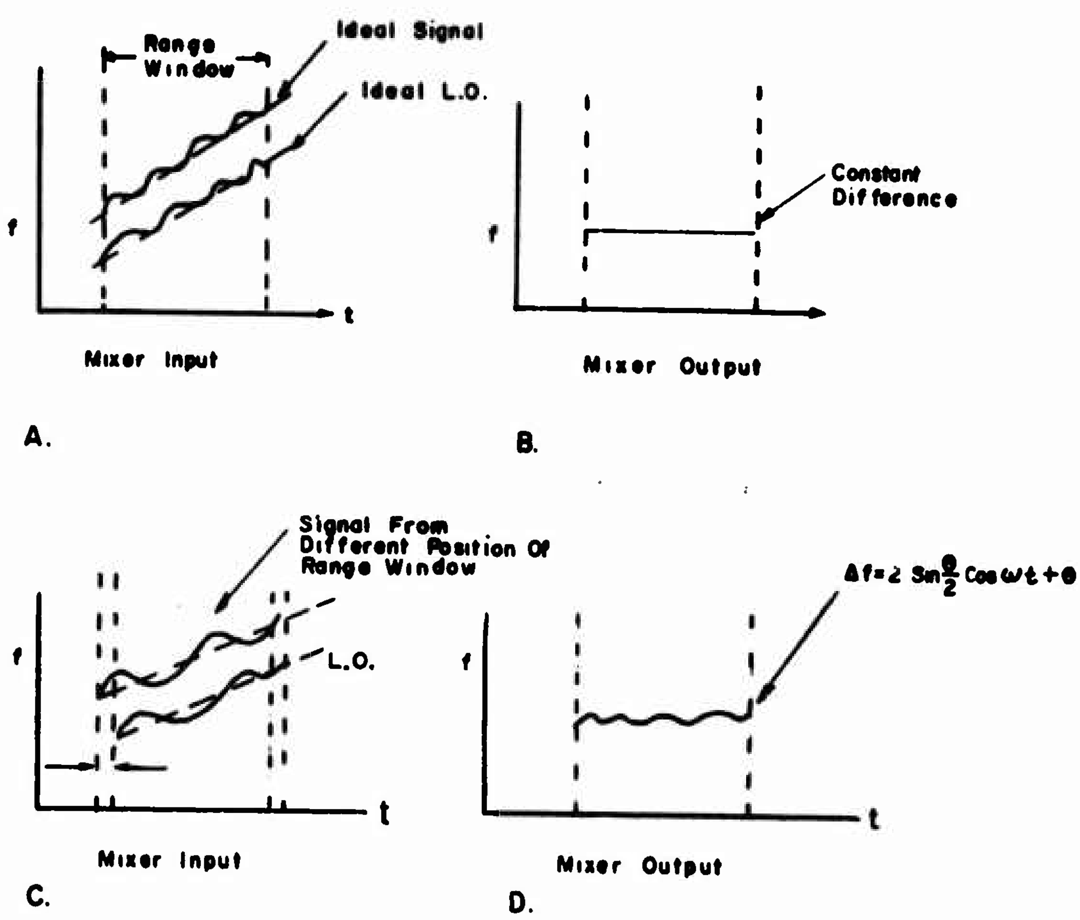


Figure 22. Effects of Nonlinearities

let

$$\Delta f = k_2 \left[\sin 2\pi f_m t - \sin 2\pi f_m (t - \Delta t)\right] \tag{5}$$

represent the deviation of frequency about the nominal IF signal. Using the trigonometric relationship for the difference of two sine waves the mixer output difference frequency error is

$$\Delta f = 2\, k_2 \sin \frac{2\pi\ f_m\ \Delta t}{2}\ \mathrm{Cos}\ 2\pi\ f_m t - \pi f_m\ \Delta t \tag{6}$$

as shown in Figure 22D.

It can be seen from the above that K_2, the peak frequency deviation; f_m, the modulating frequency; and Δt, the time shift caused by targets not centered in the range window, should be kept small. As the window length increases, this effect is more pronounced. The nonlinearities in the active ramp are to be held to 0.005% of center frequency (2.4 GHz).

A second effect of the range window on the accuracies for the pulsed L.O. type radar is the diminishing correlation energy as the target is moved away from the center of the window. That is, the mixer output is maximum for the time limited signal and L.O. when they precisely coincide. The output becomes progressively less as the target moves away from the center. Since 0.1 μsec (100 ft) does not represent much slippage of a 40-μsec pulse, most of the degradation will probably be due to the post L.O. filter. It appears feasible to maintain an amplitude drop off of 0.5 db across the window.

The range sidelobes of the signal should be specified for Mode II as well as Mode III. Since Mode II is an intermediate mode, -30 db sidelobes should be sufficient. In Mode II, the target complex will appear essentially as a point target to be tracked. The -30 db sidelobe level will minimize the possibility of tracking on a sidelobe which can be in error by 40 μsec, or approximately three miles. Mode III sidelobes, as discussed previously, are to be -35 db. The -35 db appears to be reasonable and achievable within the state of the art, which will permit simultaneous observation of specular and edge scattering phenomena within a dynamic range of at least 35 db.

Range resolution is, among other things, a function of the relative amplitude of the targets in question. A good way to insure low sidelobes is to specify a feasible resolution requirement on two point targets of different amplitudes. Referring to curves published by Bell Telephone Laboratories[19], it appears that two point targets separated by four feet and differing in amplitude by 34 db should be resolvable. The range resolution should be constant across the entire range window and is considered only for targets of equal velocity.

Doppler shifts due to the high velocities of satellites must be considered. Since there are several modes of operation, the effects on each mode must also be considered. At the operating center frequency of the radar (3350 MHz), the doppler shift can be as much as 100 Kc. For the Phase II angle track mode, the bandwidth of the 40 μsec pulse is 25 Kc and therefore the signal could be out (at least partially) of the bandpass of the track receivers. The orbital data sheets list the expected satellite velocities. These velocities will have to be translated to doppler frequency for the Mode I angle track. The Mode II bandwidth of 2.5 MHz can also have a programmed doppler shift, however, it is probably not necessary. Mode II is for intermediate ranging, and the change in range due to doppler shift will be a small fraction of the intended range resolution. The two effects of doppler on the wide band pulse (Mode III) must be considered. This is because the target velocity causes a change in slope of the frequency modulated waveform, as well as a shift in center frequency, thereby causing an error in absolute range. The latter is of little importance since absolute range is not of prime importance and amounts to only several feet. The change in slope causes a mismatch in the pulse compression filter. The additional change in slope can be estimated as follows. Let a zero doppler signal be represented by

$$F(t) - \text{rect } t \left[\sin \left(\omega_c - \frac{\omega_\beta}{2} + \frac{\omega_\beta t}{T}\right) t\right] \tag{7}$$

where

ω = center frequency

ω_β = bandwidth of ramp

T = duration of ramp (zero doppler)

and

$$F(\omega) = F^{-1} [F(t)] \tag{8}$$

since the doppler shifted spectrum is equivalent to a change in scale. Then the doppler shifted spectrum equals

$$F\left(\frac{\omega}{k}\right) \text{ where } k = \left(1 \pm \frac{2v}{c}\right) \tag{9}$$

and conservation of energy is ignored.

The Fourier transform of the doppler shifted spectrum is

$$F^{1}\left[F\left(\frac{\omega}{k}\right)\right] = \left|\frac{1}{k}\right| F(kt) \tag{10}$$

Therefore, the doppler time function is

$$F_d(t) = \left|\frac{1}{k}\right| \text{ rect } kt \left[\sin\left(\omega_c - \frac{\omega_\beta}{2} + \frac{\omega_\beta kt}{T}\right) kt\right] \tag{11}$$

The new center frequency and bandwidth are $\omega_c k$ and $\omega_\beta k$ respectively, and the new slope is

$$\frac{\omega_\beta k^2}{T} = \frac{\omega_\beta}{T}\left(1 \pm \frac{2v}{c}\right)^2$$

A good approximation of the slope is given by

$$\frac{\omega_\beta}{T}\left(1 \pm \frac{4v}{c} + \frac{4v^2}{c^2}\right) \simeq \frac{\omega_\beta}{T}\left(1 \pm \frac{4v}{c}\right) \tag{12}$$

Therefore, it is seen that the ramp slope changed by $\pm 2\,(2v/c)$ times the bandwidth. For a 16 percent bandwidth

$$\omega_\beta = 0.16\omega_c$$

the ramp change in time is

$$2\left(\frac{2v}{c}\right)(0.16\omega_c) = [2\,(0.16)]\left[\frac{2v\omega_c}{c}\right] \tag{13}$$

Thus, the slope mismatch due to doppler effects is equal to twice the signal fractional bandwidth times the signal center frequency doppler shift. For a 100 kHz center frequency doppler shift, this corresponds to 32 kHz.

It is expected that the pulse compression filter will have a sensitivity of 25 kHz per foot, i.e., frequencies differing by 25 kHz will be relatively delayed by a time equivalent to one foot in range. It is seen that the 32 kHz change in slope would cause extreme smearing of the resolved pulse. The system waveform generator should compensate for this problem and reduce this effect to, say 1% or 250 Hertz, which corresponds to a radial velocity on the order of 200 feet per second.

In addition to the doppler, the waveform should be compensated for the dispersive effects of the waveguide. Approximately 60 feet (round trip) of RG-48/4 waveguide or equivalent will be used. The first order approximation to the waveguide dispersion is considered to be linear with frequency. If the delay is linear with frequency, the effect would be to stretch the pulse. It seems reasonable that this effect could be virtually eliminated by making the input pulse to the waveguide shorter, thus making the waveguide part of the overall pulse expansion system.

The signal processor (or transmitter) must be able to keep the amplitude and phase characteristic of the transmitted spectrum within tolerable limits. These limits are roughly those which permit relatively poor sidelobe response yet they are good enough to allow for correction with a transversal equalizer. In addition, unless the transversal equalizer is quickly variable, the amplitude and phase response must be stable. The philosophy is taken since there is considerable uncertainty in what the tube characteristics will be. A linear phase is selected as a reference which can be thought of as representing the non-distorting time delay through the transmitter. This is represented in Figure 23.

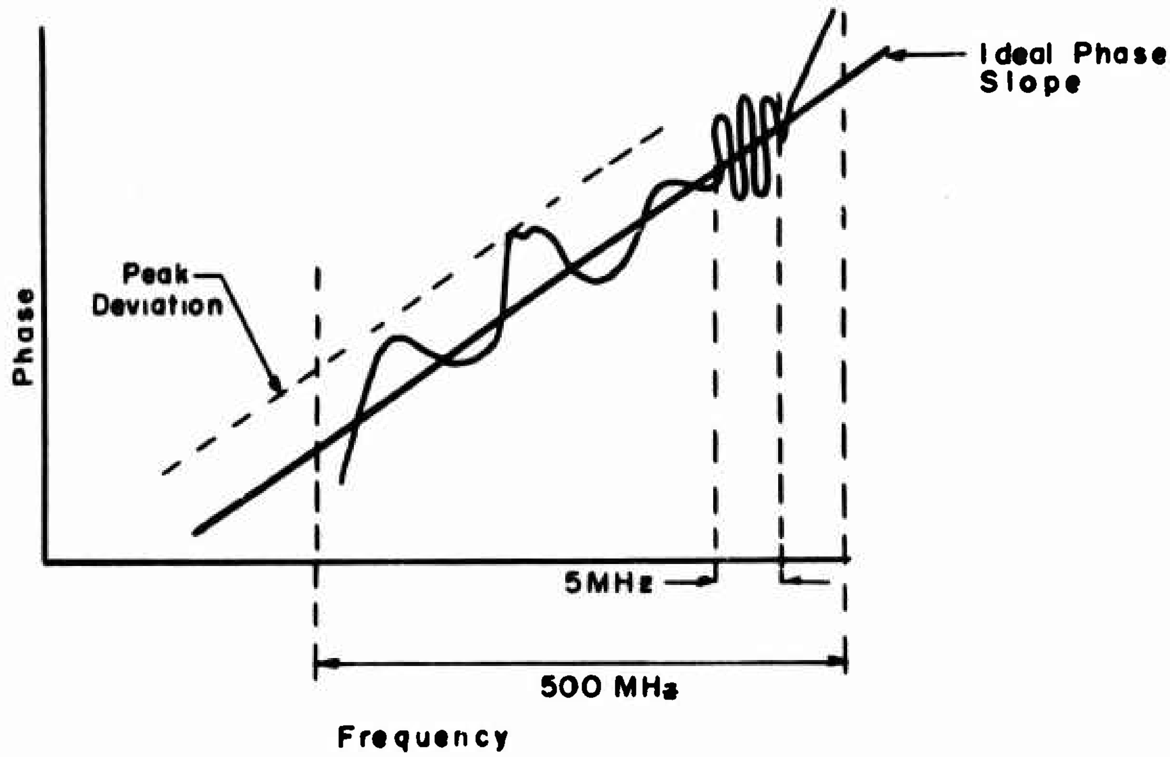


Figure 23. Transmitter Phase Response

The line can be further specified by having the integral of the actual phase response about the line equal to zero. The deviations of the phase response about the line can then be thought of in terms of paired-echo theory. It has been shown that the distortion sidelobes resulting from phase and amplitude modulation in active components have an approximate one-to-one correspondence with paired-echo distortion sidelobes for the case of a linear FM waveform. From paired-echo theory the period of a sine wave determines the position of a sidelobe, whereas its (sine wave) amplitude determines the amplitude of the sidelobe. If the peak errors are small, a fairly reliable

estimate of the echoes contributed by transmitter response can be obtained. Since the peak phase error response may result from a random-like curve, the peak of the sine wave contributing to the random peak will necessarily be less. It is also important to note that since only a 100-foot range window is being used, the sidelobes developed outside the window need not be considered.

If τ is 200 x 10^{-9} seconds, then for a 100-foot window $1/\tau$ is 5 MHz per Hertz.

This results in only considering phase-vs-frequency fluctuations less than one Hertz per 5 MHz of bandwidth since more rapid variations fall outside the window. The specification therefore should include the concept of integrating the phase-vs-frequency curve about the reference line in 5 MHz intervals, thus virtually removing the rapid variations.

To permit proper design of the signal processing subsystem, a transmitter phase of 20° peak deviation from linear is chosen. If the transmitter is in itself correctable to, say $\pm 5^{\circ}$, then the transversal equalizer may be used to further correct the overall system response. It is necessary, however, to have the pulse-to-pulse phase corrections matched to within $\pm 2^{\circ}$ or the transversal equalizer correction will not be effective over the total run of approximately 20 minutes.

Dynamic range is an important consideration for this signal processor. A 45 db dynamic range was chosen as a reasonable achievable linear instantaneous range. An additional 40-db dynamic range is required for range variation of target. The dynamic range should be specified in terms of intermodulation cross-products. That is, the linear portion of the input amplifier (tunnel diode), which is not expected to achieve more than 45 db, should not result in intermodulation products that are comparable to the sidelobe response. Selective attenuators preceding the amplifier should be used to keep the signal in the linear portion.

The decision to switch in the attenuators may be manual; however, automatic decisions and switching may be incorporated, if practical. In any event, the AGC will be constant across the range window and the amount of attenuation will be recorded as auxiliary data. It should also be noted that the limited dynamic range of the tunnel diodes will be somewhat alleviated by the use of a pulse compression waveform.

Channel matching is required to maintain an accurate relationship between the wideband cross polarized channels. The amplitude and phase match should be within 0.5 and $\pm 0.5^{\circ}$, respectively. This should enable subsequent analysis to determine some of the polarization characteristics of the signal. Future expansion to dual transmit polarization is also anticipated.

A low power signal processor checkout is expected to be performed. This should include 35 feet of waveguide to demonstrate the compensation capability. In addition, it may be desirable to put in high VSWRs to demonstrate the performance of the transversal equalizer. It is anticipated that a means to provide a closed loop checkout system will be available. For this purpose, a bi-directional coupler is to be included in the high power waveguide. This will allow simulation of a point source target and is probably the best way to evaluate overall system performance.

An accurate range tracker is required which will be capable of tracking on the high resolution data. This difficult problem will require a more detailed study. Much of the problem is due to the way one chooses to process the radar data. Specifically, the question to be answered is - which point within the 100-foot range window should and can be tracked.

Signal-to-noise ratio, scintillation, acceleration and sub-nanosecond timing are all problems to be considered in arriving at a solution. Four range tracking modes will be considered.

- Prominent point
- Electrical center of gravity
- Pre-programmed (center of mass)
- Mid-point

The prominent point method can be used for short periods of time dependent on the relative motion of the scatterers. If the target is rotating with respect to the radar, the prominent point in all probability will exist for a short time with subsequent jumping of the tracker to a new point. Also, a problem exists in relation to where the point lies within the range window, i.e., if the prominent point is on the end of the 100-foot range window, most of the target could be gated out if the window started at that point. Smoothing the total response from the target complex and determining its electrical center is another possibility. It is evident, however, that this point also wanders with target rotation. Mid-point tracking implies the range extent of the target will be utilized to find a center reference. An advantage of this method is that the target will remain in the window. However, this method also causes a jitter in the absolute radar range. One method which may possibly be developed for use is the pre-computed range track. It is an open loop type which depends on the precision obtained in determining the orbital parameters of the satellite on previous passes. This method would essentially track the center of mass of a target complex and should be the smoothest point, since it is the reference for the satellite mechanics and rotation is referenced about it. Here, too, however, it is conceivable that some targets will not lie completely in the window. Range track acquisition should occur on a signal-to-noise ratio of 3 db at maximum range. Range track accuracy, however, is dependent on the signal-to-noise ratio and should be specified in absence of noise. It appears that 0.1 foot jitter is feasible. Memory track should also be included for signal fading or mode transition.

In conclusion, it appears that the feasibility of the system is dependent to a large extent on the signal processor. Specifically, the signal processor must have the capability to compensate for other system component errors and smearing caused by propagation and doppler effects. The degree to which time varying errors can be eliminated is somewhat uncertain; however, static errors can be minimized in the system envisioned. Additionally, the degree to which range tracking can be achieved and the technique used will substantially affect the method of data processing. Satisfactory performance within the mission of the Signal Processing Test Facility is expected, however.

F. DATA HANDLING

A suitable data recording and playback capability is absolutely essential to permit subsequent detailed computer analysis of the enormous quantities of data which will be gathered on each satellite pass by the radar. As the following analysis will show, such recording would be extremely difficult, if not impossible, to accomplish when utilizing analog recording techniques. Fortunately, the signal exhibits a fairly low duty cycle, permitting the use of time expansion/bandwidth compression techniques, thus making digital recording feasible. Such recording involves real time A/D conversion and buffer storage of the short burst signal and time expanded playback of this digitized signal from a suitable buffer memory at a rate which state-of-the-art digital recorders can accept. Auxiliary data, such as frame identification, time, angle, angle rates, range and range rates, etc., are also digitally encoded and recorded in a time-multiplexed manner. Non-real time re-recording of this data in a more compatible complex format is then accomplished, when desirable, for subsequent detailed analysis, as mentioned previously.

The requirements which necessitated the utilization of such techniques are tabulated below:

- Instantaneous amplitude dynamic range: 45 db.
- Absolute amplitude accuracy over the specified dynamic range aperture: ± 1/4 db.
- Input signal bandwidth: 2.5 to 3.5 MH.
- Absolute phase accuracy: ± 1 degree out of 360 degrees.
- Input signal format:
    - Phase I: Four separate 20-microsecond duration signals, either two each amplitude and two each phase, or two each in-phase (I) and quadrature phase (Q) signals; may be serially multiplexed, if desirable.
    - Phase II: Essentially the same as Phase I signal, except that the time duration of each signal is now 40 microseconds.

It should be noted that the input signal format may consist of either separate amplitude and phase signals or in-phase and quadrature phase (I and Q) signals, whichever is more suitable. I and Q signals exhibit the very desirable potential advantage that the recording bandwidth per channel can theoretically be made half the coherent input signal bandwidth. This is not true of the amplitude or phase detected spectra, however, because such detection involves the process of autocorrelation. The theoretical advantage of utilizing I and Q recording may be difficult to realize in this application, however, because of the increased accuracy required of the individual recording functions to achieve the specified overall accuracy, as compared to that required using separate amplitude/phase recording. Even if the potential bandwidth reduction capability is not realized, however, the use of I and Q channel recording may exhibit other advantages. An important one is that a 360-degree capability wideband phase detector will not be required; such a phase detector is not easily fabricated to meet the stringent requirements of this application. Greater RMS phase measurement accuracy may also be possible, especially when the S/N ratios are high.

For the purposes of this exercise, full bandwidth recording has been assumed equivalent to that required by an amplitude/phase signal format. Similar accuracies and dynamic ranges have also been assumed. If it is later determined that I- and Q-signal recording is more desirable, the overall recording requirements are sufficiently similar to permit almost direct substitution.

1. Analog Recording Considerations

Initially, it would appear that standard analog recording techniques might be adequate to meet the aforementioned requirements. Analog machines with 45-db dynamic range and 2.5-MHz bandwidth are presently available. The limiting parameter, however, turns out to be the ± 1/4 db absolute amplitude requirement.

One-quarter db accuracy is another method of stating ± 3 percent amplitude accuracy. Thus, the sum of any erroneous signals contributing to this error, regardless of their source, must not exceed ± 3 percent of the absolute signal amplitude at that moment, wherever it appears within the specified 45 db dynamic range. In decibels, however, one signal which is 3 percent of another is approximately 30.5 db less than the other signal. Thus, the true signal-to-error signal ratio, such errors having been contributed by whatever cause, must equal or exceed 30.5 db in order to achieve the specified 1/4 db accuracy.

This 30.5 db desired signal-to-undesired error ratio requirement practically eliminates all analog recording techniques. The dynamic range of most analog recorders is specified from a one-to-one minimum S/N ratio to the maximum within the allowable distortion tolerances. A 45-db dynamic range under such circumstances would in reality be unusable over at least the first 30.5 db range due to the recorder-noise-induced error alone. This appears to be a valid assumption, if we consider this noise to have a bipolar Gaussian distribution. In this instance, the RMS value of such noise is equal to the maximum allowable three percent of the desired signal, to achieve the required 1/4 db absolute accuracy. Since the RMS value of Gaussian noise is also equal to the first sigma value, one can further conclude that the probability of exceeding the RMS value, and thus the 1/4 db absolute accuracy, is approximately 32 percent. If the RMS value of the noise is made equal to 1.5 percent of the desired signal, the probability of exceeding the three percent level is reduced to approximately five percent. The minimum required S/N ratio now increases to 36.5 db, however.

An analog recorder exhibiting a conventionally-specified 45-db dynamic range, therefore, would only exhibit a usable dynamic range of less than 15 db and ± 1/4 db absolute accuracy. This is obviously unacceptable for this application, and other techniques must therefore be found.

The only potentially-feasible, but not too practical, analog magnetic recording technique to achieve the desired stringent characteristics would be to cascade approximately three conventional 45-db dynamic range recording channels. This technique has been successfully implemented in the past with some degree of success. A three-channel recorder would be desirable, with each channel handling only 15 db of the signal at the required 30 db or greater S/N ratio through suitable attenuators and channel switching.

Unfortunately, such three-channel recorders would be extremely expensive to procure and operate, since they are not standard at this bandwidth. In addition, the difficulties of achieving the equalization necessary to maintain the required ± 1/4 db accuracy appear to place such a requirement beyond the present state of the art. Numerous and uncontrollable variations in the magnetic tape quality and record/playback functions, as well as other negative factors, resulted in the decision not to pursue such an approach at this time where assured performance is required.

Other analog techniques, primarily optical recording upon film, might also be utilized, but these, too, unfortunately, exhibit undesirable drawbacks and/or limitations. Film is being considered as a possible emergency back-up capability, however, because it is theoretically capable of providing the required characteristics, and has also been previously implemented in practical configurations. The chief disadvantages of its use are the extremely slow readout rates and the possibility of errors due to film shrinkage, if not properly compensated. Readout is presently accomplished by optical scanners at such a slow rate that even a relatively modest amount of data would require excessively long readout time, making it extremely undesirable except under emergency conditions. The necessity to continuously recalibrate in order to compensate for film distortion slows the process even more. If linear processing is used, a resolution of one part in 5700 total linear amplitude levels would be required. This is obviously not easily accomplished.

In all the aforementioned techniques, only linear recording has been considered. Would the application of logarithmic recording techniques have any advantages? Analysis indicates that they would indeed be useful if the record/playback characteristics are sufficiently linear and stable. The basic technique involves pre-conversion of the linear analog signal to a logarithmic function, and the equating of the recorder's error signals (due to noise, instability, non-linearity, etc.) to the variation in the logarithmic function commensurate with the allowable error.

Consider the application of such a technique to the previously described 45-db linear analog recorder. The maximum RMS signal to RMS "noise" amplitude ratio is expressed by the antilog of (45 db/20), which is 178 to 1. Each resolution cell will now be directly related to a constant percentage power increase in a logarithmic system. Let this value be one-quarter decibel per resolution cell. The total dynamic range will then be 178 cells x 1/4 db per cell, or 44.5 db. This just meets our requirement. Unfortunately, the linear dynamic gain/amplitude characteristics of an analog recorder can easily vary uncontrollably by 1 db or more. This, in turn, represents a linear error of approximately 11 percent in amplitude which transforms to a minimum error of 11 percent of 44.5 db in the aforementioned logarithmic configuration, or 4.9 db error. Such an error is obviously untenable in the present application. Thus, this particular technique, although somewhat promising, does not appear to offer the required accuracy and reliability for this application without further development.

2. Digital Recording Considerations

This brings us to the possible utilization of digital techniques with appropriate A/D and D/A conversion. The following material describes techniques which surmount these previously mentioned problems and discusses additional problems which are encountered and their solutions.

The most severe problem area encountered in the digital approach centers about the A/D converter. Unfortunately, the state of the art of such converters does not permit one to achieve the desired instrumentation capability directly or easily. Five basic, inter-related parameters complicate the problem. These are: (a) the analog sampling rate, (b) the sampling aperture characteristics, (c) analog-to-binary conversion rate, (d) the total number of available binary levels, and (e) the overall accuracy.

The A/D sampling/conversion rate is extremely important since the digital recording of analog signals necessitates the use of sampling techniques, and thus must meet the necessary sampling criteria. An effective rate of 10 MHz, approximately three to four times the nominally anticipated signal bandwidth, was selected after careful study, even though it severely strains the capability of existing state-of-the-art A/D converters. This might at first appear to be excessive, since sampling theory indicates the minimum rate to avoid ambiguous response need not exceed twice that of the maximum spectral components of importance. Practical considerations, however, indicate a higher rate to be very desirable, if not necessary.

First, the effective spectrum of the signal to be recorded may ultimately exceed the previously stated 2.5 MHz since the signal processor design is not finalized. Second, some bandwidth expansion is also anticipated as the result of taking only a limited number of samples preceding and following the main signal block. Third, the use of a sampling rate which exceeds the minimum value previously stated simplifies the construction of a practical sampled signal-to-analog signal reconstruction filter. The reason for this lies in the difficulty of achieving the required rate of change from bandpass rejection in the filter as the sampling rate approaches the minimally allowable two to one ratio, and this allowable aperture approaches zero. A higher sampling rate, therefore, can greatly simplify the design of a practical filter for this application.

As previously stated, however, the instrumentation problem is further complicated by the unfortunate fact that 10 MHz sampling rate A/D converters are not state of the art in the bit capacity and accuracy required to process the signal as required in a linear manner. Five-MHz rate, eight-bit resolution and potentially eight-bit accuracy converters are presently available commercially. Since eight binary "bits" can express 256 levels, this corresponds to a dynamic range of 48 db in a linear system, assuming a one-to-one desired signal to sampling error signal ratio at the lower limit. Since less than ± 3 percent amplitude error can be tolerated to achieve the required ± 1/4 db accuracy, however, at least 33 of the available equivalued levels must be used to establish the sampling error at the minimum anticipated signal level. For higher amplitude signal levels, the fixed increment amplitude sampling error represents a diminishing percentage of the overall signal, and its effect, therefore, diminishes correspondingly, becoming less important.

In order to establish the accuracy of one part in 33, however, using binary coding, slightly more than five bits are required. Thus, if a total of only eight bits are available, assuming linear processing and equivalued levels, only three bits are available to provide a maximum dynamic range of 18 db, commensurate with the desired 1/4 db accuracy. Such a limited dynamic range is not acceptable in this instance.

A possible method of circumventing this difficulty is to provide an amplitude sampling aperture whose percent error is a fixed value relative to the absolute level being sampled. Fewer discrete levels are consequently required to achieve both the desired accuracy and dynamic range for any given sample. One can easily demonstrate that the amplitude aperture is logarithmically related to the absolute sampling level with such a technique.

The logarithmic characteristic can be achieved by at least three basic means. The first is to develop an A/D converter with logarithmically spaced amplitude apertures. Although this is theoretically feasible, the technique is not presently available, requiring additional development. This uncertainty, plus the state-of-the-art availability of other techniques, makes such an approach appear to be undesirable at this time.

The more practical approach presently appears to be the utilization of state-of-the-art analog techniques to derive the logarithmic function. One method involves the use of non-linear devices to derive the exact, or nearly exact function. Another method approximates the logarithmic function with successive linear functions, utilizing suitable amplitude switching. At present, the successive approximation technique appears to be the most feasible for the dynamic range and accuracy required, but this does not necessarily rule out the use of carefully selected non-linear devices. To achieve the $\pm$ 1/4 db absolute accuracy requirement, each linear segment can cover an amplitude range of approximately - 83% to + 120% of a nominal value. This corresponds to approximately 3 db dynamic range per section. A minimum of at least 15 sections would therefore be required to cover the total of 45 db dynamic range requirement.

The aforementioned calculations are somewhat optimistic, however, because they attribute all of the allowable $\pm$ 1/4 db accuracy to the logging function. In practice, the A/D converter will also introduce appreciable errors, primarily as the result of finite amplitude sampling apertures. If one more realistically divides the errors equally between the logging and A/D conversion process, the logging error now reduces to $\pm$ 1/8 db error. The number of linear stages required to synthesize a 45 db dynamic range logarithmic characteristic now doubles to a total of 30 instead of 15 for $\pm$ 1/4 db accuracy. The practicability of such a design is somewhat questionable.

When the analog input signal is logarithmically converted before encoding, each equivalue linear amplitude sampling aperture subsequently represents a fixed logarithmic increment. Since an eight-bit binary code represents 256 linear amplitude increments, the 45-db dynamic range can be uniformly divided across the total aperture, resulting in an accuracy of (45 db/256 levels) or 0.176 db per level. If the error-per-sample does not exceed one level, the corresponding maximum error due to this effect cannot exceed 0.176 db.

The fact that the amplitude sampling error in a logarithmic system of the type described remains a fixed percentage of the absolute signal level being sampled raises the question of whether the reconstructed analog signal may be adversely affected. The analysis of this potential degradation can be simplified by the assumption that the actual sampling function is the sum of two synchronous sampling functions; one is exact, the other contains only the error. The effect of the error function can then be quantitatively evaluated by the summation of the corresponding impulse response of the reconstruction filter to these error samples superimposed upon its response to the exact function.

Linear expansion of the logarithmically compressed signal must be accomplished before the reconstruction process, however. Since the logarithmically-compressed signal is already digitized, a computer operation appears to be the most logical expansion method. Unfortunately, it will do nothing to correct any errors associated with the signal.

It is well known that the optimum time weighting function for each sample of a sampled signal to affect distortionless reconstruction exhibits a Sin X/X characteristic having its zeroes at multiples of the sampling rate. If such an optimum function is used, it is immediately apparent that the reconstructed signal amplitude should be equivalent to the sample amplitudes at those periods of time coinciding with each sample. If the previously mentioned sampling error of $\pm$ 1/4 db is not exceeded, the reconstructed signal accuracy should therefore be acceptable at least at these periods of time. One is not only interested in accuracy at a point, however. The next very important question, therefore, centers about the accuracy of the reconstructed signal at all times other than the sampling times.

Since the signal is reconstructed by the process of summing the time-weighted, time displaced samples, it is apparent that temporal weighting sidelobes generated by errors may cause reconstructed signal errors at other periods of time. One can further conclude that the absolute signal error is proportional to the absolute signal magnitude for a fixed relative error. As a result, the absolute error associated with the temporal sidelobes of the reconstruction weighting function is similarly related. This leads to the immediate conclusion that the weighting sidelobes of the strong error signals pose the greatest threat to weak signals. This problem sounds very familiar, since pulse compression signal processing gives rise to the same undesirable phenomena.

For example, if one assumes the temporal resolution capability of the time-expanded pulse compression signal to be 0.6 microsecond, and the sampling rate is ten MHz the attenuation characteristic of the optimum Sin X/X reconstruction function will be:

$$A = \left(\frac{1}{\pi\ ft}\right)^2 = \left(\frac{1}{6\pi}\right)^2 = \frac{1}{354}$$

corresponding to -25.5 db. Since signals of 45-db dynamic range are of interest, it is quite apparent that strong signal sidelobes can far exceed the desired small signal in such instances.

Fortunately, however, it is only the error component which is of interest in this instance. If this error does not exceed $\pm$ 1/4 db, its relative magnitude will not exceed $\pm$ 3 percent of the true signal. This, in turn, represents an error signal which is at least -30.5 db less than the true signal. When this error value is additionally weighted by the temporal sidelobes, the resultant error level coincident with the adjacent signal resolution cell is now (-25.5 db) + (-30.5 db) = -56.0 db.

This overall induced error level is good, but not good enough to provide the desired 45-db dynamic range. A weaker, adjacent signal must be stronger than (-56.0 db + 30.5 db) or -25.5 db relative to the stronger signal in order to prevent a 1/4 db strong signal error from inducing a similar error in the smaller signal as a result of the temporal sidelobes associated with the signal reconstruction function.

This, in reality, potentially limits the adjacent signal dynamic range to 25.5 db unless corrective measures can be taken. Fortunately, if the weighting is accomplished by a computer, a suitable correction should be possible to effectively negate this source of error. The success of this technique is predicated upon the assumption that (a) the sampled values represent the true signal values (excluding basic logging and A/D conversion errors) from which a correction signal can be derived and (b) the correction signal is essentially a sinusoid.

The latter assumption should be reasonably correct, since, for weighting filter $f\tau$ values of six as previously calculated, the attenuation varies less than 2.5 percent as $f\tau$ varies from 6 to 7, for example. (This parameter could also be corrected, however, if it is later considered to be necessary.) This, in effect, makes the temporal sidelobe response for $f\tau$ values of six or greater essentially a sinusoid.

Computer weighting, is therefore considered to be essential to obtain the desired overall reconstruction accuracy over the specified 45 db dynamic range.

The use of a computer is also dictated for additional reasons. An exceedingly important one is the ability to synthesize a perfect reconstruction filter characteristic without amplitude or phase distortion. Such distortion, almost impossible to escape if a practical analog filter were utilized, could introduce paired echo distortion of sufficient magnitude to seriously affect the aforementioned dynamic range/absolute accuracy capability.

In addition, a computer provides the means to correct adverse gating modulation characteristics which may otherwise generate serious error signals during reconstruction. This potentially serious problem will be minimized by (a) additional sampling for ten percent of the normal gate width immediately preceding and proceeding the gate and (b) the introduction of additional correction/weighting signals into the computer generated signal reconstruction process.

The previous brief discussion has demonstrated that the sampled signal can be reconstructed into its original analog form with the desired dynamic range and absolute accuracy if the sampled values exhibit similar characteristics. It has also been pointed out that the two main sources of sampled signal error in a practical system are the logging and A/D conversion functions. The A/D converter requirements and problems are now discussed in more detail below.

Theoretically, the sampling function should approximate an impulse function, but this is impossible. The most usual implementation of such circuits is a gated capacitor holding circuit with appropriate charging feedback. As the sample time decreases, especially in the nanosecond region, it becomes increasingly more difficult to implement practical circuits. Gating impedances ultimately approach impossibly low "on" values and infinitely high "off" impedances in order to achieve the required capacitor charging accuracy. For one percent sampling accuracy before integration, nanosecond apertures or less are desirable for this application.

High speed A/D conversion as described herein requires the use of an analog signal sample and hold function. This can be a source of serious error.

If one can assume that the gating function associated with the sampling aperture approaches a square wave function with appropriate transfer and non-transfer impedances and the gate is followed by an appropriate integrator, such as a low pass filter, it may be possible to achieve the desired absolute sampling accuracy without the requirement for excessively small sampling apertures. A theoretically perfect sampling aperture of approximately 10 nanoseconds followed by a suitable integrator should be adequate to achieve a one percent amplitude sampling accuracy. Twenty-five nanosecond apertures with this degree of accuracy are available today on commercial 5 MHz sampling rate, eight-bit A/D converters, but, unfortunately, this accuracy cannot be achieved with signal bandwidths much beyond 1 MHz. Nanosecond and sub-nanosecond samplers are also available, primarily as part of sampling oscilloscopes, but their accuracy may exceed five percent. It is believed that, for use with a suitable integrator, a feasible sampling aperture will consist of a five to ten-nanosecond effective width, when sampling 2.5 MHz bandwidth analog signals.

The error introduced by the analog sampling function can be analyzed in a manner similar to the analysis of quantizing level error. Since a three percent amplitude error represents 1/4 decibel error, the achievement of a one percent absolute sampling error should keep the degradation from this particular source well within limits. Since the two sampling functions are cascaded (sample and hold, and A/D conversion), their corresponding errors can be cumulative, although in practice, they may tend to complement and counteract one another. The design objective is to keep the cascaded peak error of the two sampling functions below the + three percent value commensurate with the $\pm$ 1/4 db overall accuracy.

The previous material has concentrated on those areas which may cause the greatest difficulties. The rest of the system will now be described more generally, since it primarily embodies proven digital techniques.

As previously stated, the required 5 MHz, eight-bit A/D converters are commercially available. These utilize signal delay techniques to serially accomplish the digitizing action, beginning with the most significant bit. Modified input sample and hold circuits will most likely be required, and additional modifications, such as improved reference signal stabilization, may be required to achieve the desired accuracy. Two of these units will be time multiplexed to provide the required equivalent sampling rate of 10 MHz. Their individual output will appear on appropriate buffer registers supplied with the basic instrument.

If separate amplitude and phase signals are to be recorded from the cross-polarized receiver channels, four sub-blocks of data are involved. The data length of each sub-block will be a minimum of 20 microseconds with the 250 MHz system and 40 microseconds with the 500 MHz system. The most likely form of processing is to serially multiplex these analog signals through suitable delay media, leaving an appropriate space between each sub-block. If we allow a five-microsecond buffer on each end of each data sub-block of the 250 MHz bandwidth signal for weighting control, and separate each sub-block plus its buffer zone by an additional 10 microseconds, the result will be a 150-microsecond time duration overall data block. Similarly, the 500 MHz signal will have an overall data block of 230 microseconds.

The digital buffer memory/time expander must accept the digitally encoded data contained in the data blocks at an effective 10 MHz rate. This poses an instrumentation problem, since the nominal read-in rate of various practical memory devices does not match this rate. Except for active flip-flop type circuitry, the average read-in rate of most other, less expensive, and small memory devices is approximately one microsecond per bit, although a large number of bits may be recorded in parallel. The possible exception to this may be the thin film technique which may exhibit a faster rate, if properly implemented.

Since a memory of at least 2000, eight-bit words will be required to handle only the high resolution data, the use of active shift register techniques, while feasible, does not appear to be very feasible in cost, complexity, reliability or size. Cores, on the other hand, are reasonably inexpensive and reliable. Two approaches can be used independently or together to meet the required read-in rate requirements. The first approach is to sequentially multiplex 10 separate eight-bit word core memories, which although feasible, is not very practical. The second approach is to reformat the sampled output from the A/D converters into larger words, which can be read out in parallel format at a slower rate into a larger size memory. To reduce the 10 MHz A/D converter output rate to one MHz would require the formatting of at least an 80-bit word. This is impractical. If the memory is split in half, however, with each taking a 40-bit word at the one MHz rate, a realizable capability should be achievable.

Thus, a practical method of implementation would involve the use of two multiplexed buffer core memories, each capable of storing 200 40-bit words. Operating at a 5-MHz read-in rate, an active register would most likely be used to reformat the output of each A/D converter to generate the one MHz rate, 40-bit word. A small increase in word size would also be desirable to permit parity checking and should pose no additional problems. All operations would be suitably synchronized and multiplexed by the use of standard techniques.

A somewhat more expensive, but more practical, buffer memory might be implemented using four separate core memories. This would reduce the required input read-in rate, after multiplexing to 0.5 MHz which is well within the state of the art. Approximately 170 24-bit words would be required, not including any extra capability for parity checking, which is quite reasonable.

Provisions must then be made to read out the buffered data at a time expanded, reduced bit rate compatible with the digital recording capability. Since the basic radar PRF is 70 per second, the overall interpulse time is approximately 14,286 microseconds. Of this period, up to 150 microseconds will be occupied with the gathering of

approximately 1200 eight-bit samples at the 10 MHz rate, for the 250 MHz radar signal bandwidth. Correspondingly, for the 500 MHz radar signal bandwidth case, up to 230 microseconds will be occupied with gathering its 2000 samples. If we round off the data gathering period to 286 microseconds, we are left with 14 milliseconds per pulse repetition period in which to play back the digitally encoded signal from the buffer memory at a reduced rate suitable for recording upon magnetic tape.

Assuming each sampled eight-bit signal is consecutively read out of the buffer for direct recording, the required readout/record rate for the 250-MHz radar signal will be 70,000 words per second, and 140,000 words-per-second for the longer duration, 500-MHz radar signal. The lower rate can easily be recorded by commercially available digital recorders, which, at a packing density of 800 bits per inch and a tape speed of 120 inches per second, are capable of recording 96,000 bits per second per recording channel. Recording of the other signal using this method would require the multiplexing of two recorders or at least double the previously stated tape speed of a single machine. The use of two machines, although not impossible, would generate serious synchronization problems. Higher tape speed machines are feasible, but are not off-the-shelf items.

A logical approach appears to be the simultaneous recording of two eight-bit words, on a 16-parallel channel machine after suitable reformatting. This lowers the required recording data rates to one half of their former values, namely, 35 kilobits per second and 70 kilobits per second. Such recorders are commercially available. Since the two eight-bit words will potentially consume all of the available channels transversely on tape, parity checking might practically be accomplished in the longitudinal dimensional. A suitable shift register will be necessary to provide the proper reformatting and interim buffer storage after core memory readout so that it can be properly recorded.

Since important auxiliary data must also be recorded, along with the desired signal for each PRF (i.e., frame number, range, angle, radiated power, noise figure, AGC setting, etc.), the recording rate and total density will of necessity exceed the values quoted above for the signal alone. In addition to tape identification and satellite catalog number, the individual requirements are tabulated below.

AUXILIARY DATA REQUIREMENTS

FUNCTION	INCREMENT	TOTAL UNITS	NO. BITS
Frame Identity	1 Frame	126,000 Max Frames	17
Absolute Range	1 ft. accuracy over 400 miles	2.112 million	21
Azimuth Angle	0.0014°	$0\text{-}360^{\circ}$	18
Elevation Angle	0.0014°	$0\text{-}90^{\circ}$	16
Radiated Power	0.1 db	6 db	6
Receiver Input Atten.	five steps	40 db	3
Receiver Noise Fig.	0.1 db	10 db	7
Absolute Time	1 usec	24 hrs (86.5 billion usecs)	37
Day of Year	one day	365 (+1)	9

The aforementioned data is quite significant in importance and quantity, but, unfortunately, all but the last four should be recorded again with every new frame of signal data. (The necessity for continuously recording such data as noise figure and transmitter power with every frame may be questionable and can stand further examination.) This adds complexity to the recording process, but, fortunately, does not require much space on the recording tape relative to the signal data. As a result, the required recording rate including this data is not much more than for the signal data alone. Preliminary calculations indicate, for example, that the frame rate auxiliary data could be formatted to occupy 12 longitudinal cells or lines per frame of a 16-parallel channel tape. Since the signal data will occupy 1000 longitudinal cells per frame, this represents an increase of only 1.2 percent. Even if we assumed a boundary limit requirement of 20 longitudinal lines per frame, this additional two-percent requirement in the number of lines would

increase the maximum recording rate per channel from 70 kilobits per second to 71.4 kilobits per second, a very small and tolerable increase.

Suitable buffer registers will be required to store and feed the auxiliary data to the tape recorder, as required. Whether there may be advantages in multiplexing various buffering processes has not yet been evaluated. If the data can be serially multiplexed in the core buffer memory/time expander play back, it would appear that the multiplexing might be both feasible and desirable. Additional control circuitry will be necessary, of course,to provide all of the required functions implicit with the auxiliary data recording function.

It is apparent at this time that the recording function, while not easy, is within the present state of the art. Sufficient reserve capability should exist to permit required parity checking and reliable operation. Two suitably multiplexed machines will be required to permit continuous recording, since the 120 inches-per-second rate or 10 feet-per-second, only permits approximately two and a half minutes of total recording per 1500 feet. The operator will of necessity find a majority of his time taken up with the changing of tape reels during actual satellite tracking and data gathering runs. Suitable automatic equipment will also be required to bring the standby machine up to speed and on-line as required.

The final item of discussion concerns whether the data, as recorded, can be directly used without reformatting or additional processing for computer reduction and analysis. This problem, although not serious, has yet to be resolved. Present plans include on site, off line re-recording the signal in IBM format. This will require another machine and suitable reformatting and control capability. In addition, since the relative packing density per unit length of magnetic tape will have been more than halved because of the reduced number of parallel channels, it will also require at least twice the original length of tape to accomplish. Where 9000 feet would suffice for 15 minutes of the 16-channel recording, over 18,000 feet will be required for the reformatted computer compatible tape as now contemplated. This can obviously consume great quantities of tape, if data is gathered on a number of satellite passages and a permanent or semi-permanent record is desired.

In conclusion, a technique has been described which should be capable of recording the data derived by an extremely high resolution radar with a high degree of accuracy over an extended dynamic range. Such a capability unfortunately is not commercially available today as a packaged unit, but the basic techniques required to fabricate it fortunately do exist. Although implementation problems are involved, there is no reason to believe that they cannot be surmounted. Perhaps with the development and fabrication of this initial unit, coupled with a growing requirement for such a recording capability, industry will meet the challenge to make similar, if not improved, capability units commercially available.

G. SERVO SYSTEM

The controlled movement of the radar antenna, as required in both azimuth and elevation for satellite acquisition and tracking, is presently accomplished by a servo system employing electronically controlled hydraulic motors. Two such motors are used in a back-to-back or opposing torque configuration to provide anti-backlash movement in each dimension. Primary hydraulic power is provided by an electrically driven, constant pressure, variable displacement pump operating in conjunction with various accumulators to provide an essentially constant pressure, variable rate, low loss source of hydraulic fluid. Extensive fluid filtering is utilized to remove foreign contaminants. Fluid flow into the motors is controlled by magnetically energized valves driven by appropriately compensated electronic amplifiers to provide a type II servo system. Negative velocity feedback is used to generate a synthetic damping constant for stability. Various analog and digital angle readout devices are also provided to generate appropriate reference and control signals. A number of driving or error input signals may be used, including those derived from synchros, digital equipment, and monopulse antenna techniques. Manual "slewing" is also possible.

The ability of the servo system to achieve the required pointing angles, rates, accelerations and accuracies described herein and in Appendix B is absolutely essential to permit the accomplishment of the mission. The ultimate objective of the servo system investigation, therefore, is to quantitatively evaluate the present capability and, if deficient in terms of the new requirements, to delineate a design configuration which is acceptable. For a number of reasons, however, the investigation to date has been quite limited.

The qualitative conclusions of this brief study can be summarized as follows:

(a) The basic electro-hydraulic drive system as now implemented can achieve the required maximum and minimum angular velocities and accelerations in open loop operation.

(b) Additional investigation is necessary to adequately determine the quantitative transfer functions of the basic electro-hydraulic drive system in order to definitely conclude whether or not the desired characteristics can be achieved in closed loop operation.

(c) Manufacturer's design data indicates that the strength of the present bearings and gears may not be adequate to provide optimum life when subjected to the anticipated velocities and accelerations. Additional investigation is therefore necessary.

(d) The basic closed loop servo control electronics will most likely require redesign to provide a more stable, type II zero velocity error capability. This is considered to be a minor modification.

(e) The digital controller and its ancillary equipment must be redesigned or retrofitted. The objective is to increase its readout rate by a sufficient factor to negate the present sampling rate errors which will otherwise make it unacceptable for use in the planned configuration. A modest modification is anticipated.

The desired servo directing characteristics can be summarized as follows for the specific functions indicated.

a. Long Range Acquisition Involving "Cross Track" Error Scanning only at a Fixed Range.

The basic scanning method will be the use of a circular or one dimensional sinusoidal scan. The maximum angular deviation in one dimension will approximate 0.2 degrees. To achieve a high data replenishment rate, a maximum sinusoidally varying acceleration of four degrees per second squared is anticipated in either dimension with a maximum angular velocity of approximately 0.60 degrees per second.

b. Medium Range Acquisition Involving Error Volume Scanning.

The basic scanning method will be the use of a computer directed scan which successively searches along the anticipated satellite track for different assumed values of cross track error with each scan. In order to maximize the data replenishment rate, an angular velocity of up to five or six degrees per second is desired during the constant velocity portions of the scan, with a maximum of four degrees per second per second acceleration during a sinusoidal scan reverse phase. Tracking accuracy should be within at least ±0.05 degree of that commanded in both dimensions during this phase of operation except during scan reversal, where up to ± 0.5 degrees error should be permissible in the maximum acceleration dimension.

c. Computer Directed or Closed Loop Monopulse Tracking.

Except during those high zenith passages of the satellite requiring rapid azimuthal traverse and re-acquisition, the maximum anticipated angular velocities required for satellite tracking will approximate four to five degrees per second with maximum accelerations up to 0.4 degree per second per second in both dimensions. Normally, however, these requirements will be much less. The minimum "smooth" angular velocity should be 0.001 degree per second.

The rapid azimuth traverse will utilize a linear five to six degree per second maximum angular rate with sinusoidal acceleration/deceleration at the extremities of this rapid traverse limited to a maximum of four degrees per second per second as in the volumetric scan mode. The tracking error should not exceed + 0.02 degree of boresight except during acquisition, this value representing approximately a + 12 percent beamwidth error.

One factor which has unfortunately limited the preliminary investigation has been the unavailability of original design data. The most serious result of this situation is the inability to ascribe a realistic transfer function to the electric valve/hydraulic motor complex. Other important parameters relevant to the mechanical characteristics of the antenna mount are also in doubt. A number of experimental tests have been conducted in an attempt to determine the missing data, but in many instances the lack of suitable instrumentation has severely limited the useful quantitative data which could be gained. Nevertheless, tne total data which is now available serves to establish certain boundary conditions, which, in turn, make possible tentative conclusions of a reasonably optimistic nature.

One of the most basic determinations is whether the electro-hydraulic drive system has sufficient torque and power to achieve the required angular acceleration and velocity characteristics. Fortunately, the requirement for maximum of either acceleration or velocity usually occurs when the other is a minimum. Since the power expended during acceleration is proportional to the product of acceleration and velocity, the net instantaneous driving horsepower requirement is minimized.

Experimental open loop measurements indicate the maximum steady state velocity to be approximately 5.5 to 6.0 degrees per second and the initial step response acceleration to be approximately 4.5 degrees per second squared. These peak values of one when the other is essentially zero are more than adequate; however, their relationship at intermediate values when both occur simultaneously is equally important. As previously mentioned, the lack of suitable instrumentation unfortunately prevented the conduct of worthwhile quantitative experiments to gather such information.

Appropriately corrected theoretical calculations provide a fairly good indication of this intermediate performance. The basic high pressure hydraulic pump is rated at 18 horsepower output at 2200 psi. This in turn is driven by a 40-horsepower electric induction motor. Each of the hydraulic motor complexes in both azimuth and elevation is indicated to be capable of approximately 175 ft lb of torque with a pressure differential of 2200 psi. This potential torque is tnen multiplied by 640 in the drive gear train to a maximum potential of 1.12×10^5 ft lb to position the antenna in either dimension.

The torque of the motors is resisted by at least three basic forces common to all mechanical rotating systems. These are directly proportional to acceleration, velocity, and a fixed constant as expressed by the following relationship.

$$J\left(\frac{d^2\theta}{dt^2}\right) + D\left(\frac{d\theta}{dt}\right) + K_F = \text{Applied Torque}$$

where

J = Ft lbs per radian per second squared

= moment of inertia

D = Ft lbs per radians per second

= damping constant

K_F = Ft lbs running friction

The value of J, the moment of inertia, has been determined by very crude experiments to be approximately one million ft lbs per radian per second squared in both dimensions. D, the damping constant, has never been directly measured or calculated, but from wind gust effects, a tentative value of 40,000 ft lb to 50,000 ft lb per radian per second has been estimated. Running friction, K_F, was also crudely measured to be approximately 10,000 ft lb. With $(d\theta/dt)$ not exceeding six degrees per second, it is immediately apparent that the damping factor term will be negligible. Expressed in terms of ft lbs per degrees per second, D equals 700 to 875. At breakaway, when $(d\theta/dt)$ is very small (i.e., $d\theta/dt < 0.1$ degree/second) maximum motor torque can be generated since little available fluid driving horsepower is actually required and maximum pressure and flow is available. As the angular velocity increases, the power absorbed by the accelerative and running friction forces also increases.

$$\text{Accelerative Power} = T(a) \times \left(\frac{d\theta}{dt}\right)$$

$$= J\left(\frac{d^2\theta}{dt^2}\right) \times \left(\frac{d\theta}{dt}\right)$$

$$\text{Friction Power} = T(f) \times \left(\frac{d\theta}{dt}\right)$$

$$= K_F \left(\frac{d\theta}{dt}\right)$$

Input Power = Acceleration Power + Friction Power

Again, the effects of damping are believed to be too small to be of any consequence. Sooner or later, however, the limited available fluid drive horsepower also limits the available acceleration at various angular velocities. Assuming the maximum available fluid horsepower to be 15, a curve, Figure 24, has been plotted showing available acceleration in degrees per second squared vs angular rate in degrees per second. The dotted line also shows the theoretical acceleration which could be obtained if the motors were appropriately variable displacement or coupled through an appropriate torque converter to utilize the full 15 horsepower at all angular velocities. The friction horsepower is also plotted as a function of angular velocity to indicate its power absorbing effect. Gear loss has been lumped under friction loss for these calculations.

The calculations indicate that maximum motor acceleration of approximately 5.8 degrees per second squared can theoretically be maintained for all angular velocities up to 4.2 degrees per second. If the maximum available fluid horsepower is assumed to be 15, the acceleration then becomes total power limited for angular velocities above 4.2 degrees per second dropping to a value of 3.9 degrees per second squared at six degrees per second angular velocity. It is quite apparent that these values definitely exceed the previously stated maximum acceleration/velocity requirements to achieve the desired operation. Even if the actual values of moment of inertia are 25% higher than assumed herein, calculations indicate that the net acceleration should still be more than adequate.

There is some question, however, whether the higher values of acceleration and velocity desired for the new operation may possibly exceed the stress/wear limitations of the gears and bearings. The only pertinent design information which is presently available indicates that the operating limits which were assumed for design purposes were somewhat lower. This data may not be accurate, requiring a thorough investigation of this factor.

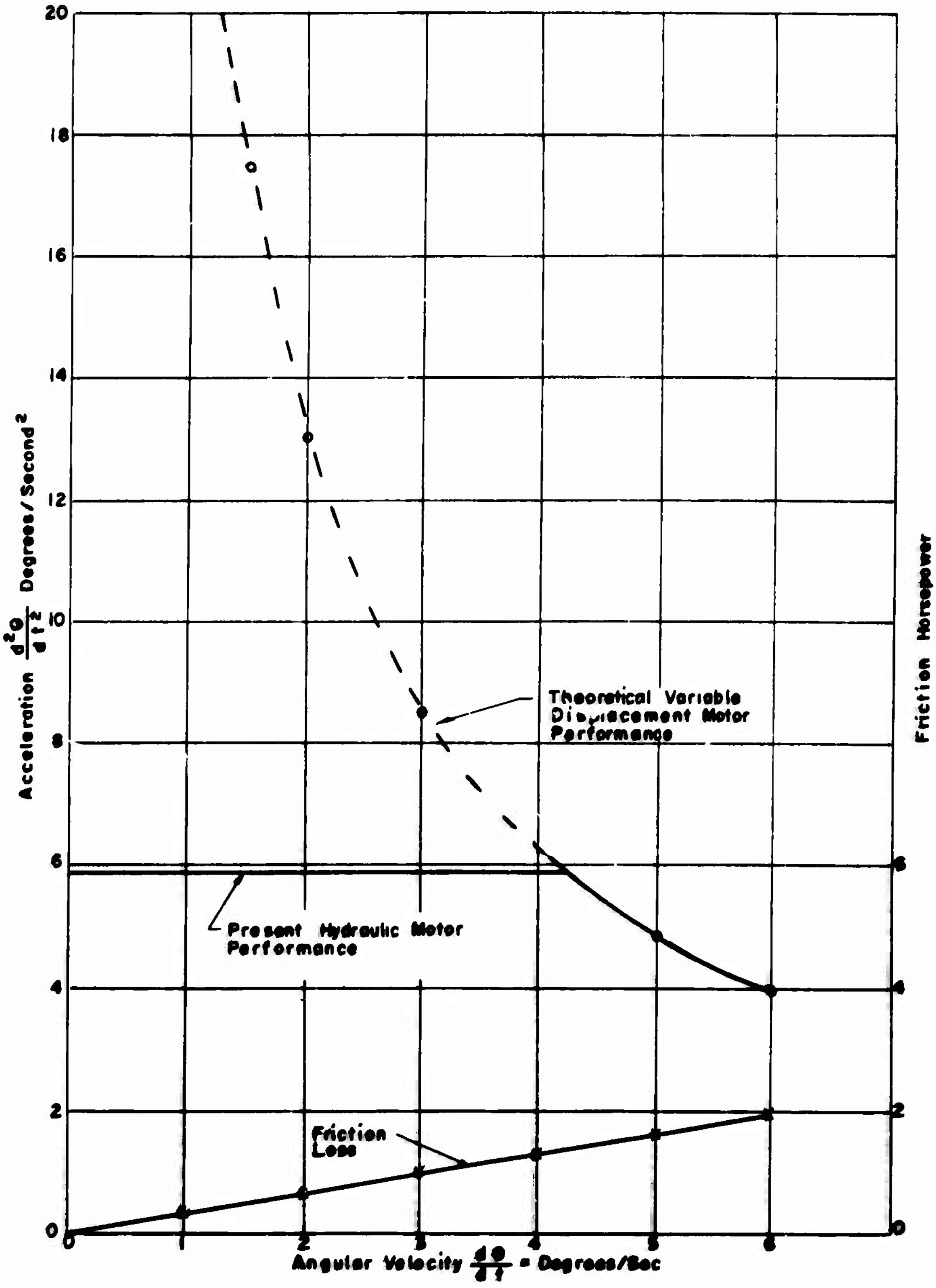


Figure 24. Hydraulic System Performance

Although the maximum velocity/acceleration capabilities of the basic hydraulic complex should be adequate, this, unfortunately, does not also permit the conclusion that the present complete configuration is also adequate or that it can be made adequate to provide the desired dynamic tracking capability with its associated accuracy and stability. Additional testing and analysis will also be necessary to make such a determination, and such an effort is presently underway.

As stated earlier, the most difficult aspect of this effort appears to be a reasonably accurate determination of the combined hydraulic motor/valve transfer function. The valve function is apparently the most nonlinear element involved (except when the pressure sensitive relief valves are activated), its flow being a function of both valve opening and pressure. Figure 25 illustrates the quantity/pressure flow relationship of a typical hydraulic valve designed to control one motor for various metering displacements. The pressure function is actually the differential pressure measured between the hydraulic motor driving port and the exhaust port. The flow is thus controlled by two metering valves which are in series.

The characteristics of the valve function are further complicated in the present configuration due to a somewhat unorthodox use of a single valve to drive two opposing motors to achieve reduced backlash. As a result, the exhaust of each motor is ported to a common sump. Instead of the previously mentioned series flow of the hydraulic fluid through the two sections of the valve when used with one motor, fluid now follows two separate paths. One path involves flow from the high pressure inlet through one section of the control valve and the driving motor to the sump. The other motor is driven backward, pumping fluid from the sump through the other valve section and back to the sump. This latter pumping action may contribute significant negative torque, especially during overspeed, deceleration conditions. Determination of its contribution to the overall transfer function is therefore extremely important. The valve characteristics, illustrated in Figure 25, must obviously be modified to incorporate this different technique of operation.

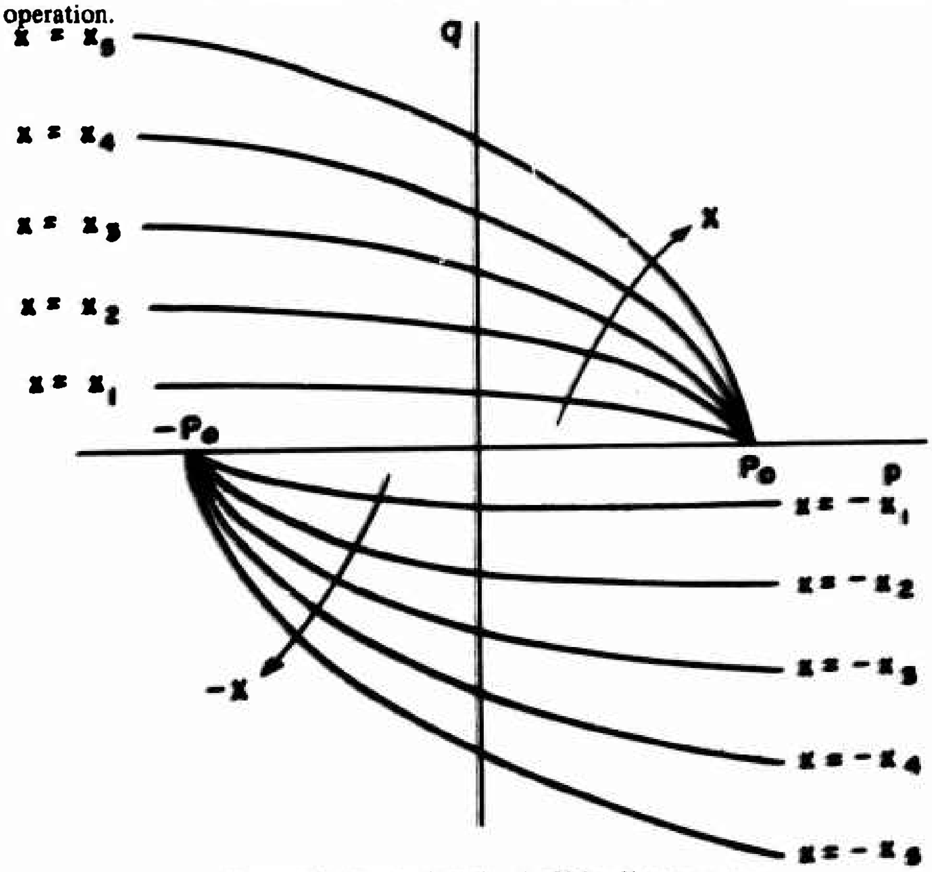


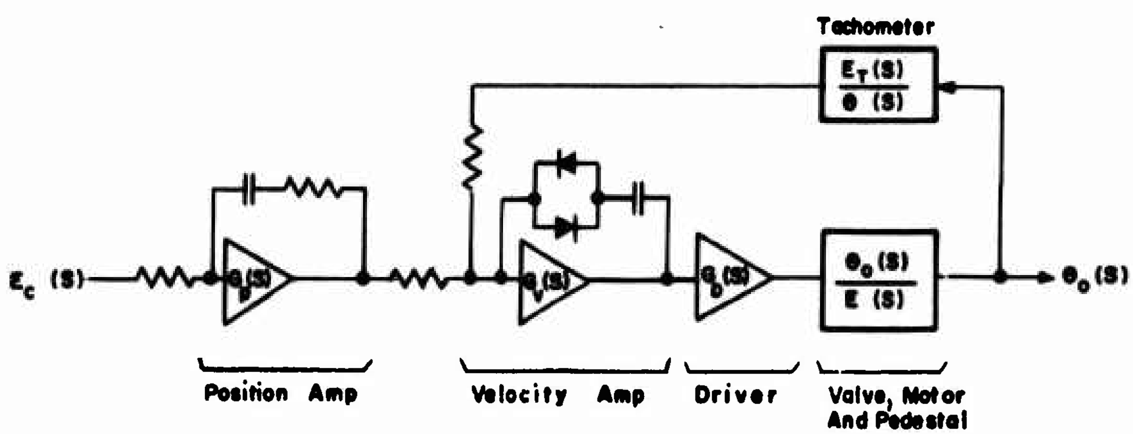
Figure 25. Typical Hydraulic Valve Characteristics

Various instrumentation and test methods can be suggested to measure the desired transfer characteristics of the valve/motor complex without removing it from the antenna pedestal. The difficulties of providing suitable instrumentation to achieve the desired dynamic accuracy, however, has led to a tentative conclusion that the basic hydraulic drive complex should be physically removed from the pedestal and evaluated upon a suitably instrumented test stand. This course of action will receive further consideration to determine its practicality.

It is almost axiomatic that at least a type II servo system will be required in order to achieve the minute errors which have been specified for achievement during constant velocity and acceleration tracking. The basic type II system should be capable of zero error constant velocity tracking; however, it will develop an error during acceleration. The required acceleration values may be sufficiently small, however, that the resultant error can be kept within tolerable limits.

Figure 26 is a block diagram of the present servo system when operating with position error derived from a monopulse antenna. It incorporates both a position and velocity amplifier, with tachometer derived velocity feedback around the velocity amplifier, and monopulse position error around both amplifiers. Both the velocity and position amplifiers utilize basic operational amplifiers incorporating additional feedback to provide a specific transfer characteristic.

Analysis of the position amplifier indicates that its output is composed of two functions, one which is proportional to the input signal and the other which is proportional to the time integral of the input signal. The equivalent transfer function can be expressed as $G_p(S) = A_1 + B_1/S$ and is achieved by an R-C feedback signal around the otherwise linear operational amplifier. The feedback function around the velocity amplifier, on the other hand, generates two mutually exclusive transfer characteristics. When the feedback voltage is insufficient to fire the zener diodes in the feedback loop, the loop is essentially open providing high linear gain properties. $G_v(S)$ then equals D for all



$$G_p(S) = \left[A + \frac{B}{S}\right]$$

$$G_D(S) = 1$$

$$\frac{\theta_o(S)}{E(S)} = \frac{K}{S(S+\alpha)}$$

$$G_v(S) = \begin{bmatrix} D & \text{For } -E_z < E_v < +E_z \\ \frac{F}{S} & \text{For } E_v > |E_z| \end{bmatrix}$$

$$\frac{E_T(S)}{\theta(S)} = SK_T$$

Figure 26. Servo System Transfer Response

values of feedback voltage across the zener diodes which are less than their conduction potential of approximately three volts. When the feedback voltage is sufficient to equal or exceed the zener conduction voltage, the feedback capacitor is connected and the transfer function becomes that of an integrator. In this case $G_V(S)$ equals (F/S).

A rather simplified analysis then indicates that the present configuration exhibits two different servo characteristics depending upon the status of the nonlinear velocity amplifier transfer function. In order to show this, let us assume that the hydraulic valve/motor combination exhibits the following characteristics:

- Motor Torque $\propto$ Differential Pressure, P.
- Fluid Flow through Motor $\propto$ Motor Velocity, $(\frac{d\theta}{dt})$.
- Fluid Leakage Around Motor, etc., $\propto$ Differential Pressure, P.
- Fluid Compressibility = 0.
- Fluid Flow through Valve $\propto$ Control Voltage, E_c.

The total hydraulic fluid flow through the valve and motor can then be expressed as

$$q_{valve} = K_v E_c = [q_{motor} + q_{leakage}]$$

$$= [K_m (\frac{d\theta}{dt}) + K_1 P]$$

Fluid compressibility has been ignored, although in practice its effect may tend to cause unstable operation. Since motor torque is proportional to pressure, the equation above can be rewritten as follows.

$$T_M = K_o [q_{valve} - q_{motor}]$$

$$= [K_2 E_c - K_3 (\frac{d\theta}{dt})]$$

This expression is quite similar to the electric motor torque equation in which the field excitation is held constant and the control function is applied to the armature, as seen below.

$$T_{em} = [K_4 E_c - K_5 (\frac{d\theta}{dt})]$$

If we then equate the hydraulic torque of the motor to the torques due to inertial, velocity, and friction effects and solve for the transfer function $\theta(S)/E_c(S)$, we find

$$\frac{\theta(S)}{E_c(S)} = \frac{K_6}{S(S+a)}$$

Combining this function with the velocity amplifier transfer response "D" (with no feedback through the zener diode loop) and adding the tachometer derived rate feedback, results in an overall transfer response measured at the input terminals of the velocity amplifier of

$$\frac{\theta(S)}{E_v(S)} = \frac{K_8}{S(S + K_9)}$$

Further combining this function with the transfer function of the position amplifier results in the following overall transfer function

$$\frac{\theta(S)}{E_p(S)} = \frac{K_{10}(S + K_{11})}{S^2(S + K_7)}$$

This represents a type II servo system which would theoretically be capable of zero error under constant velocity tracking conditions.

When the zener diodes conduct and the velocity amplifier transfer response becomes F/S, the new transfer response relating angular output versus velocity amplifier input with tachometer rate feedback becomes

$$\frac{\theta(S)}{E_v(S)} = \frac{K_{12}}{S(S^2 + SK_{13} + K_{14})}$$

Further combining this with the position amplifier transfer response results in the following overall response

$$\frac{\theta(S)}{E_p(S)} = \frac{K_{15}(S + K_{16})}{S^2(S^2 + SK_{13} + K_{14})}$$

This still represents a type II servo response with additional poles as a result of the integration. Whether it will be stable, and capable of achieving the desired dynamic tracking accuracies with or without additional modification, remains to be more accurately determined.

It is interesting to note the action which the two states of the velocity amplifier have on the velocity feedback function. With small errors and corresponding non-conduction of the zener feedback diodes, the velocity feedback function is linearly amplified to serve as a large synthetic damping constant. As soon as the error increases to the point of conduction of the zener feedback diodes, the resultant modification of the velocity amplifier into an integrator effectively converts the velocity feedback signal to one which is directly proportional to angular displacement of the output shaft. This and other facets of the servo electronics must receive further study and evaluation.

Turning to the digital control portion of the servo, it is quite evident that modification may be necessary in order to make it compatible with the new requirements. As stated in Appendix B this system comprises a digital computer and appropriate ancillary instrumentation to smoothly drive the antenna along pre-computed look angles for the tracking of satellites of the Echo class where only low values of angular velocity and acceleration are involved. Since the computer is presently incapable of on-line look angle calculations from the basic satellite ephemeris data, the required calculations must be accomplished beforehand. The computer then operates as a readout and interpolative device in real time, as well as recording auxiliary data. Five sets of pointing angles are presently computed for satellite arrival times of 0, ± 1/2 minute and ± one minute. The purpose of this lead/lag data is to correct for

the pointing errors which such arrival time delays introduce because of the earth's rotation. The method of second order interpolation which is then used to provide a more accurate correction supposedly provides an infinite correction capability between these extremes of delay.

The basic deficiency of the present digital configuration is its slow readout rate of once-per-second. This is too slow to provide the unambiguous control signal bandwidth which is required to achieve the anticipated servo system characteristics while under this form of control. Since the overall analog servo system bandwidth will most likely exceed 5 Hz, a minimum digital sampling rate requirement of at least 10 Hz is anticipated. Operating it in sub-synchronism with 1/2 the PRF of 70 Hz might be advantageous.

Unfortunately, the read-in/read-out rate of the present PB-250 digital computer cannot be increased beyond twice a second without major modifications. The lack of a suitable fast access memory is the basic cause of this restriction. Studies have indicated, however, that a major modification or replacement may be avoided by the use of an auxiliary function which can synthetically generate a suitable wideband control function utilizing the present once-per-second readout rate. One method of accomplishing this is to utilize an auxiliary function which can essentially expand the once-per-second data into a power series. The coefficients of this expansion function can be corrected at the slow once-per-second rate, while the interpolative readout rate can be made almost any desired value. This approach is receiving careful consideration.

Other portions of the digital control system must also be modified or replaced to achieve the desired control bandwidth. Included are the digital angle encoders, their converters and shift register memories, and the subsequent error signal filtering and weighting functions. Since the encoders utilize optical readout, increasing their readout rates basically involves an increase of the flash lamp rate. Specifications for these Baldwin encoders indicate that rates up to 100 pulses-per-second are possible with reduced lamp life, while rates of 30 to 50 pps should provide an adequate compromise between data rate and lamp life. The present conversion rate of this data from a gray code to binary code also appears to be limited. This should not present much of a problem or difficulty to remedy, however.

The digital data rates are still so slow, even at 50 pps, that little shift or storage register difficulties are anticipated. If the present instrumentation of this type proves to be inadequate, however, standard inexpensive operational units should be capable of substitution with little or no difficulty. The subsequent filtering and weighting functions will undoubtedly require complete replacement. Assuming that the sampled readout rate is sufficiently high to achieve the required unambiguous spectral response, the filtering will most likely be in the form of a low pass reconstruction filter. The phase shift characteristics of such a filter will undoubtedly be very important, and may dictate the need for a higher sampling rate in order to optimumly shape the roll-off characteristics.

In conclusion, it is evident that additional investigation is necessary in order to determine the adequacy of the present servo system to meet the new control requirements. The basic hydraulic system has been determined to be capable of achieving the maximum velocity/acceleration values, however. On the other hand, whether the required dynamic accuracies and stability are achievable with or without some modification has not yet been ascertained. Such an effort is now underway. The present configuration does embody type II servo characteristics; these are considered to be a necessity in order to reduce velocity and acceleration errors during tracking to the minimal values which will be required. The digital control portion will definitely require modification, but this appears to be of a moderate nature. Finally, mechanical stress analysis must also be performed to insure that the integrity and life of the components involved will not be excessively compromised by the new requirements.

SECTION VI
MODIFICATION PROCEDURE

A. PROJECT PLAN

The modification of the present tracking radar will be accomplished in two ways. First, detailed design and fabrication of each of the major subsystems must be accomplished. After installation and checkout of the major subsystems, system integration will be accomplished by interfacing the subsystems into a complete integrated radar.

This effort will be accomplished by RADC personnel with the assistance of the subsystem contractors and a site modification contractor. During the course of the design of the major subsystems, the interface requirements will be specified in detail to insure compatibility between each subsystem.

During equipment development the site modification will continue mainly in the area of removing unneeded equipment and wiring. In the meantime, certain problem areas are to be reviewed and solutions arrived at. Some of these problem areas are:

- Weight limit and weight distribution of antenna mounted components.
- Antenna balancing procedures.
- Performance limitations of the servo system.
- Computer interface problems.

Any decision for further modifications will be dependent on the results of these studies.

B. SUBSYSTEM EVALUATION

As each subsystem is developed, it will be subjected to design review, test and evaluation to determine its compliance with original design requirements. The test and evaluation criteria will be developed during the fabrication phase of the subsystems. Preliminary testing will be accomplished at the contractor's plant under the supervision of RADC. Final acceptance tests will be conducted at the Floyd Site after the equipment has been installed. A critique of the results will be performed by RADC to determine the true operational performance of each subsystem. After acceptance testing has been completed, the subsystem will be interfaced with other subsystems.

C. CALIBRATION AND SYSTEM INTEGRATION

At present, techniques are under investigation for developing a complete test procedure to establish the system performance and accuracy. The use of aircraft, balloons and satellites is being considered for the tracking tests.

Antenna measurements of pattern, sidelobes and boresight are to be taken using the existing boresight tower and all angular readout systems, theodolites and telescopes collimated. After these tests, the measurements will be checked against radio stars and finally on precise orbit satellite data.

Range sidelobe performance measurement will be accomplished initially on a closed loop basis where a sample of the transmitter output will be inserted into the transmission line leading from the antenna feed to the RF preamplifiers. In this way, a measure of the performance of the signal processing subsystem and transmitter is obtained. Complete system sidelobe measurement is more complicated and will require the use of a coherent repeater at a remote site or a calibrated sphere either in orbit or balloon borne.

Atmospheric distortion is not expected to require any real effort on Phase I since a pulse broadening on the order of 10 percent to 15 percent is expected, according to available data.[21] The Phase II dispersion correction will be programmed into the system as a function of elevation angle.

D. OPERATION

During FY 67, an operational test plan for the modified radar will be developed. A catalog of all orbiting objects has been obtained from NORAD and a list of all objects of interest is being drawn up. Time-of-arrival experiments will be performed on these objects using the RADC ASFIR[22] Sites so that problem areas may be uncovered relative to the programming of the oribtal tracker and of the acquisition program due to excessive deviation in time-of-arrival. Cross-track errors and their effect on the acquisition problem are also under study.

When a catalog of objects of interest has been compiled, a test plan and schedule will be drawn up for the operation phase to take place in FY 68.

Data will be recorded on magnetic tape and will include phase and amplitude for both received polarizations as well as clock track, range, doppler correction, azimuth and elevation angles and frame identity. At the completion of a satellite pass, the data will be reformatted into binary form for later reduction in a computer.

A computer program is presently being developed for use with the "Short Pulse Model Measurement Program" for data reduction and analysis.[23, 24] This program will also be used for the data reduction and analysis of the information gathered in connection with this program.

SECTION VII
SUMMARY

The results of the preliminary design study indicate that the instrumentation of the existing Floyd Test Site to provide a high resolution sensor with modest time sidelobes of approximately -20 db is feasible.

Achievement of the desired -35 db sidelobes requires a degree of stability, linearity and matching which has not been commonly obtained in the past, requiring additional investigation and effort. This investigation has served to expose these potential problem areas both for evaluative purposes and to guide the development effort. Some of the more important conclusions are as follows:

a. Antenna

When modified with the new wideband feed horn, the overall antenna response should generally be adequate assuming that severe mechanical distortion does not occur during scanning. The effect of time delay dispersion resulting from off-axis target illumination upon the temporal sidelobes of the wideband waveform has not yet been fully evaluated.

b. Transmission Lines

Phase dispersion and reflections resulting from mismatched conditions appear to be the primary sources of wideband signal distortion. Dispersion effects are generally controllable by suitable waveguide component selection and the incorporation of signal pre-distortion compensation techniques. Mismatch and the resulting reflection may be more difficult to control, but are not considered impossible to constrain acceptable limits. Fortunately, transversal equalizers will in most instances provide a means of controlling the resultant time delayed ambiguous signals generated by such mismatches.

c. Duplexer

Although ferrite duplexers have many advantages, the difficulty of providing matched phase and low insertion loss characteristics required in each of the cross polarized receiving channels precluded their use. Gaseous duplexers should be capable of meeting these requirements, but additional difficulties arise, such as effective life, recovery time, and VSWR mismatch. Effective life and recovery time are not considered to be serious problems, although they cause operational difficulties. The effects of VSWR mismatch should generally be capable of compensation by the use of transversal equalizers. Achieving the required bandwidth does not appear to be a problem.

d. Transmitter

The transmitter is comprised of the drivers, final amplifier and pulse modulator. Basic RF signal excitation is provided by the signal processor. Little difficulty is anticipated in achieving the required signal handling characteristics in the low level driver section. The required phase and amplitude characteristics of the final amplifier tube are difficult to achieve. An existing tube, the VA-145F, is being modified for Phase I operation (250 MHz) with only theoretical data presently available attesting to its ability to meet the required characteristics. Even less data is available relative to the Phase II, (500 MHz) final amplifier tube, although that which is available is encouraging. Active phase compensation techniques are being examined to provide additional phase correction capability to insure

that the transmitter will meet its required performance. An integrally related problem involves the instability of the modulator pulse, which can lead to additional phase and amplitude modulation. It is presently believed that practical pulse leveling techniques can be applied to minimize this undesired modulation.

e. Wideband Receivers

The use of tunnel diode RF pre-amplifiers provides a reasonable noise figure, minimizes amplitude and phase nonlinearity, and simplifies cross channel matching. Dynamic range does not appear to be a problem. Since wide-to-narrow band signal bandwidth compression is accomplished immediately in the first converter following the RF pre-amplifiers, the difficulty in suitably handling the wideband signal is minimized. Little difficulty is anticipated in maintaining phase and amplitude characteristics in the converter circuit, including the local oscillator de-ramping signal to be provided by the signal processor.

f. Monopulse Tracking Channels

Little difficulty is anticipated because of the relatively narrow bandwidths involved. An important design requirement, however, is stable operation over a wide dynamic range in order to provide adequate pointing error data on rapidly scintillating targets. Such a design appears to be within the state of the art.

g. Signal Processing

The feasibility of the basic pulse compression technique which utilizes actively generated linear FM and active correlation partial de-ramping with attendant information bandwidth compression has been demonstrated. Dispersive compression filters provide a time domain output for the reduced bandwidth signals.

Compressed pulse sidelobe levels of -25 db should be achievable in system operation without undue difficulty. The desired -35 db sidelobe is much more difficult to achieve and maintain in practice, but it does not appear to be impossible. A prime and difficult requirement of the signal processor is to generate a suitably linear and repeatable FM ramp for excitation and first mixer decorrelation. Various techniques now exist which can be implemented to provide the desired characteristics. The signal processor will also incorporate techniques to compensate for microwave component dispersion and amplitude and phase ripple. Modification of the ramp slope prior to transmission will compensate for quadratic phase dispersion and phase ripples while active voltage controlled attenuators are used for minor amplitude ripple correction.

h. Data Handling

The requirement to record 40-microsecond bursts of 2.5 MHz bandwidth signal with an absolute accuracy of ± three percent over a dynamic range of 45 db is not easily achieved. State-of-the-art recording techniques cannot meet all of these requirements simultaneously. A unique time expansion technique utilizing logarithmic compression of the analog input signal followed by digital conversion and recording is under development and should be capable of providing the desired performance. The main areas of difficulty center about the stability and accuracy of the logarithmic compressors and the A/D converter. Although these problems are difficult to solve, they do not appear to be insurmountable.

i. Range Acquisition and Tracking

No difficulties are anticipated when operating in the narrow band mode, even at ranges out to 1000 miles. Intermediate modes exhibiting increased range resolution should be adequate to permit final positioning of the 100-foot range gate. Unless computer-aided high resolution range tracking is utilized, however, situations are anticipated in which range gate jitter may occur as the target scintillates. It is anticipated that a special purpose computer

can perform the required smoothing and coasting nctions most economically. Precomputed predictive data may also be utilized as an additional input. The presen. .igh resolution range tracker will be a third-order digital tracking system and is expected to perform adequately for most operating conditions.

j. Angle Acquisition and Tracking

The narrow antenna beamwidth coupled with its overall limitations of inertia and structural strength practically eliminate the use of volumetric search for acquisition. The position of the satellite to be acquired must therefore be known to an accuracy of approximately $\pm$ 2.5 miles at a range of 900 miles. Present ephemeris prediction techniques should be capable of such accuracy when using suitably updated data. The use of a computer to accomplish this function is almost mandatory. Monopulse tracking could be utilized if some form of coasting function is provided during periods of scintillation induced signal loss. A small special purpose computer could easily provide this function. A more desirable technique would be the use of computer-derived look angle commands using predicted ephemeris data throughout the required tracking period with narrow band monopulse correction. The latter technique is entirely feasible and will most likely be instrumented. The adequacy of the present servo system to provide the required operation is still under investigation; the main area of concern centers about the resultant dynamic accuracy.

k. Calibration and Test

One of the main problem areas about which very little is known at this time concerns the procedures and instrumentation required to accurately determine, calibrate and maintain the desired individual techniques and overall system performance which is required. Conventional instrumentation to accomplish the desired subsystem tests is not generally available, requiring special development, some of which is presently underway. Suitable techniques for the evaluation and calibration of overall system performance have to be more thoroughly investigated. A suitable test is the prime requirement. The use of satellites, balloons, aircraft and boresight towers is under investigation, but each exhibits unique limitations and problem areas which make any individual technique less than adequate to accomplish the required tests. In all probability, the investigation may reveal that a combination of these individual techniques will be required to achieve the desired capability.

The preliminary design effort culminated with the development of system specifications in accordance with the system parameters shown in Figure 27. The complete system specifications are contained in Appendix E.

	PHASE I	PHASE II
PEAK POWER	6.5 Mw	10 Mw
AVERAGE POWER	10 Kw	20 Kw
CENTER FREQUENCY	3350 Mc	Same
MOD BANDWIDTH	250 Mc	500 Mc
PRF	70 PPS	Same
PULSE WIDTH	20 μsec	40 μsec

Figure 27. System Parameters

REFERENCES

1. HAC Pub. No. 68-97500 Instruction Sheets for Passive Satellite Research Terminal, Hughes Aircraft Company.

2. FD 60-201 Communications System AN/FRC-40, Technical Note I, 5 December 1960, Pg. 89, unpublished report by Hughes Aircraft Company.

3. RADC-TDR-62-210, AD 276 360, "Programmed Satellite Tracking," Gordon Negus, May 1962.

4. EMC-TM-65-1, "General Satellite Look Angle Program," Kalynez and Negus, February 1965.

5. RADC-TR-66-67, Orbit Differential Correction - Tracking Program,

   Vol. I, AD 633 720, Differential Correction Geocentric Orbit Computations Program

   Vol. II, AD 633 721, Preprocessor for the Differential Correction Program

   Vol. III, AD 633 628, Earth Satellite Orbit Predication Program

6. Lothes, R.N., "Floyd Radar Modification - Final Report", Syracuse University Research Corporation, unpublished report December 1965.

7. Sciambi, A.F., "The Effect of Aperture Illumination on the Circular Aperture Antenna Pattern Characteristics," Microwave Journal, August 1965.

8. Williams, William F., "High Efficiency Antenna Reflector," Microwave Journal, July 1965.

9. The Microwave Engineers Handbook and Buyers Guide, 1966, Page 170.

10. Ruze, John, "Physical Limitations on Antenna," MIT Technical Report 248, October 30, 1952.

11. Jasik, Henry, Antenna Engineers Handbook, McGraw-Hill, 1961.

12. Jansen, Paul A., "Designing Cassegrain Antennas," Microwaves, December 1962.

13. Polk, C., "Transient Behavior of Aperture Antennas," Proc. of IRE, July 1960.

14. Kinsey, Richard R., "Monopulse Difference Slope and Gain Standards," IRE Trans. on Antennas and Propagation, May 1962, Pages 343 - 344.

15. Hannan, Peter W., "Optimum Feeds for All Three Modes of Monopulse Antennas I: Theory," IRE Trans. on Antennas and Propagation, September 1961.

16. RADC-TDR-62-24 Vol. II "Synthetic Spectrum Radar" AD 346-331.

17. Goldman, S. "Frequency Analysis, Modulation and Noise" McGraw-Hill 1948.

18. Cook, Bernfeld, "RADAR Signals" Academic Press, April 1967.

19. Klauder, M.R.; Price, A.C.; Darlington, S.; "The Theory and Design of Chirp Radars" BSTJ, Vol. 39, July 1960.

20. Cook, C.E., "Transmitter Phase Modulation Errors and Pulse Compression Waveform Distortion", Microwave Journal Vol. 6, Pp. 63-69 May 1963.

21. RADC-TR-64-76 Final Report, Vol. I, "Decoy Discrimination Studies", 3 January 1964, AD 348 743.

22. RADC-TDR-62-580, "Proc. of Symposium on Pulse Compression Techniques and Applications; ASFIR - Active Swept Frequency Interferometer Radar" p. 36, AD 336 308.

23. RADC-TR-66-357, "Signal Processing Studies and Analysis", June 1966, AD 486 291.

24. RADC-TR-66-358 "Signal Processing Studies and Analysis", June 1966, AD 486 684.

APPENDIX A
SYSTEM SENSITIVITY CALCULATIONS

A satellite identification radar with high range resolution capability should preferably exhibit a long range, high signal-to-noise detection/tracking capability of extremely small echoing areas. If the useful range is too short, it may seriously degrade the capability of gathering worthwhile range vs aspect angle data because of a limited observation time.

It has been determined that the AN/FRC-40, located at Floyd, New York, can be modified through the use of state-of-the-art techniques and components to achieve a useful capability of the type desired for range in excess of four to five hundred miles. Although this value is less than optimum, such a facility can in many instances gather extremely useful satellite data of the type desired.

The choice of system parameters has not been capricious or arbitrary. A thorough systems analysis was required, since compromises were necessitated by the availability of components, state-of-the-art, techniques, etc. The sensitivity of the resultant design is calculated below.

The essential radar parameters of the Phase I equipment are as follows:

P_{av} = 10 kilowatts average

Te = 20 microseconds (transmitted pulse)

Tc = 0.4 microseconds (compressed pulse after bandwidth reduction)

PRF = 70 pulses per second

Be = 250 megahertz (modulation bandwidth)

λ = 0.1 meter RF wavelength

NF = 6 db

L = 5 db system/signal processing loss

G_T = 50 db effective

G_R = 50 db effective

σ = 0.1 square meter echoing area

KT = -204 db

A derivation of the effective gain of the Cassegrain antenna system at the new operating frequency is included in another appendix. The 50-db value used above includes anticipated losses.

Matched filter signal processing is assumed. Those practical losses which prevent achievement of the theoretical capability are covered under the 5-db overall system loss. There does not appear to be any reason why such processing should be severely degraded or unachievable. Target velocity decorrelation effects which might otherwise become a problem are assumed to have been properly compensated.

The standard signal-to-noise vs range equation for matched filter signal processing was used to obtain the desired performance data. The individual characteristics of the complex, multistep pulse compression are ignored. Their effective overall loss is summed under system loss.

The aforementioned single pulse signal-to-noise vs range equation is:

$$\frac{S}{N} = \frac{(P_{av})(G_T)(G_R)(\lambda)^2(\sigma)}{(PRF)(KT)(loss)(4\pi)^3(R)^4(NF)}$$

Solving this equation for the (S/N) ratio after signal processing for various values of range when the effective echoing area, σ, is only 0.1 square meter results in the signal-to-noise ratios tabulated in Figure A-1.

RANGE (mi)	S/N(db)
600	13.0
500	16.9
400	20.9
300	25.0
200	32.9
100	44.9

Figure A-1. Signal-to-Noise Ratio in High Resolution Mode (Based on 0.1 M^2 Target)

This ratio has also been calculated for different echoing areas and the results are graphically displayed in Figure A-2. The increased range corresponding to a fixed, minimally acceptable signal-to-noise ratio is immediately apparent.

The Mode I narrow band long pulse of at least 20 microseconds will be used for satellite acquisition and initial angle/range tracking. As a consequence, the effective target echoing area of satellites of interest is expected to be well above one square meter because of the range resolution of approximately three miles. Consulting the signal-to-noise vs echo area curves depicted in Figure A-2 shows that a one-pulse signal-to-noise ratio of 23 db can be expected at 600 miles for a one-square meter target which can be extrapolated to +11 db at 1200 miles. This should be more than adequate for the stated purpose.

Mode II operation will provide intermediate accuracy positioning of the high resolution range gate to be used in Mode III. The bandwidth of this signal is approximately 2.5 MHz corresponding to a range resolution of approximately 0.4 mile or 2100 feet. Since the size of individual satellites will not exceed this resolution, echoing areas similar to those anticipated during Mode I operation are also expected in this instance. Again, the previously calculated one-pulse signal-to-noise ratios should be adequate for this purpose.

Mode III high resolution operation should involve much smaller echoing areas because of the approximately 2-1/2 - 3-foot weighted range resolution during Phase I operation, and approximately 1-1/4 to 1-1/2-foot resolution during Phase II operation. As a consequence, the resultant reduction in the one-pulse signal-to-noise ratio at ranges of 600 miles or greater is expected to be intolerable, necessitating the prime data gathering on potential echoing areas as small as 0.1 square meter to be conducted at ranges of 300 to 350 miles or less. Figure A-2 indicates the Phase II one-pulse signal-to-noise ratio for a 0.1 square meter target to be +25 db at 350 miles range. This is sufficient to provide the required one-pulse RMS amplitude accuracy of approximately ± 1/4 (assuming no other sources of error) and similarly an RMS phase accuracy of approximately ± 2.5 degrees.

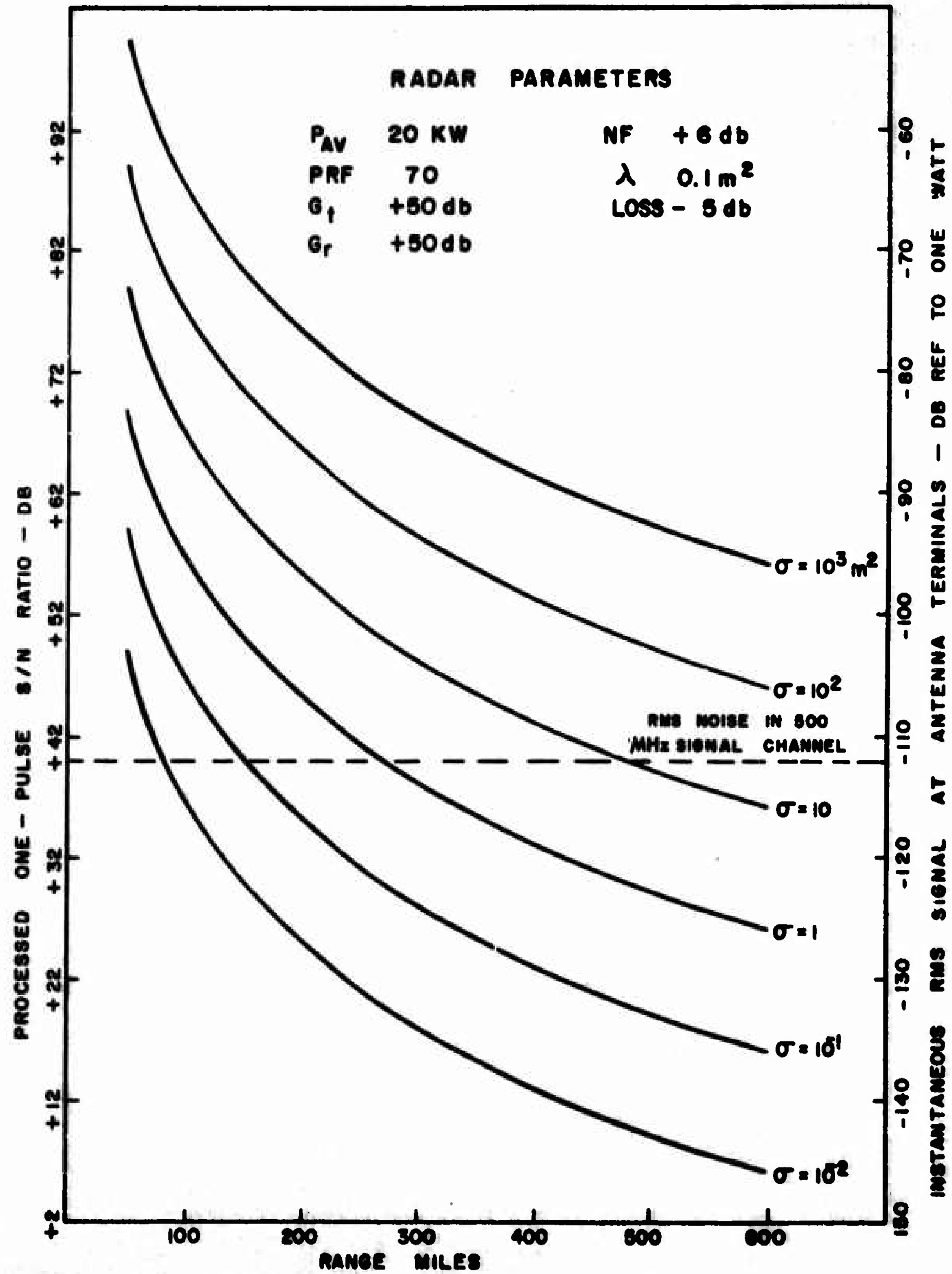


Figure A-2. Radar Performance

Successful high resolution data gathering should therefore be feasible when observing satellites having scatterers of 0.1 square meter or larger at slant ranges of at least 350 miles or less. Assuming nominal satellite heights of interest to be approximately 150 miles, those satellites exhibiting nominal or little orbital eccentricity and passing within a few hundred miles of the radar should generally be observable for at least 60 seconds of continuous high resolution data gathering. In many instances of satellites which pass closer, the useful data gathering period will be extended commensurately.

APPENDIX B
ACQUISITION AND TRACKING

A prime requirement is that the radar be capable of acquiring and tracking satellites of interest with sufficient accuracy in angle and range to permit recording of the desired high resolution data. For a number of reasons to be outlined herein, computer direction or aiding of these functions is considered to be a necessity. As presently planned, the basic orbital elements will be supplied by such agencies as NORAD/SPADATS or the Goddard Space Flight Center. Initial acquisition and tracking data will then be pre-computed using these elements. Where feasible, tracking will be initiated prior to the high resolution data gathering to provide more accurate, updated orbital information. In most cases, initial acquisition will be accomplished under open loop computer control to minimize initial uncertainty scanning requirements. Upon acquisition, the open loop computer control will be closed, utilizing actively derived angle and range data as a corrective feedback function. During the initial tracking phase before the gathering of high resolution data, this angle data will be gathered on a pulse-to-pulse basis. Later, when high resolution data is being recorded and monopulse angle data cannot be gathered simultaneously, the updating of angle information may be accomplished in a multiplexed, sampling basis. The rate may vary from once every few PRF pulses to once every ten seconds or larger, if it is demonstrated to be feasible.

For the present, the satellites of interest will include only those whose orbital elements are reasonably well known. This data will in turn greatly simplify the desired acquisition and tracking function as well as minimize the overall instrumentation requirements to accomplish it.

The satellite of interest must also meet additional requirements, if useful data is to be gathered. Synonymous with the requirement that their orbital elements be accurately known is the requirement that the parameters be fairly stable and/or well defined. This, in effect, reduces the unknown factors of interest and/or the need for frequent updating to maintain the necessary accuracy. This may eliminate certain satellites; for example, those whose physical area-to-mass ratio is so large that solar pressures and atmospheric drag cause appreciable orbital errors. Since low altitude orbits are also more seriously affected by atmospheric drag, it is anticipated that the minimum altitude of typical satellites of interest will exceed one hundred and fifty miles. Maximum altitude during high resolution data gathering is also limited by the effective data gathering range of approximately 400 miles for 0.1 square meter echo areas. The satellites' altitude and projected ground range must obviously be less than this to ensure that it will remain within the maximum useful range to permit recording of high resolution data for a reasonable period of time. With typical satellite velocities approaching five miles per second, a reasonable data recording period in this instance may only be a minute. Finally, in addition to the other requirements, low eccentricity orbits will generally be favored because the resultant rates and accelerations associated with the tracking function will tend to be minimized, commensurately reducing instrumentation requirements.

As with all very narrow single beam radars, the ability to utilize some form of volumetric scan to aid the acquisition problem is extremely limited. In this instance, the inertially related power requirements and maximum physical stress limitations are the primary restrictions preventing the achievement of high scanning rates. If these effects were not present, the maximum linear scanning rate would then be limited by the requirement for the antenna beam to remain on target for a sufficient period of time to receive the one or more target echo returns as required for detection and acquisition. The time delay of the basic PRF interval of the subject radar is commensurate with a slant range of approximately 1200 miles. Thus, it becomes the limiting interval on a one-hit per beam dwell basis for those slant ranges less than this. Since most acquisition ranges will be less than 1200 miles, PRF delay will be the predominant scan rate limiting factor as described. On a one hit per beam dwell period basis, this corresponds to a maximum linear scan rate of 23.3 degrees per second, reducing to 11.65 degrees/second maximum for a two hits per

beam dwell period requirement. Since the primary antenna beamwidth is one third of a degree circular, the maximum solid angle scanned per second with such a linear scan could not exceed 7.8 and 3.9 degrees squared, respectively, for the aforementioned one-hit and two-hit beam dwell conditions.

Practically, however, the present maximum linear scanning velocity is restricted by the aforementioned limitations to six degrees per second, and the acceleration is limited to three or four degrees per second per second. The maximum solid angle scanning rate at this linear velocity is commensurately reduced to two degrees squared per second. Whether these rates and accelerations can be materially increased without excessive physical damage to the structure, gears, and bearings has yet to be completely analyzed. Available design information, however, indicates that present gear strength and life may limit any such increase.

If small linear segments must be scanned, as may well be the case for initial acquisition, the maximum acceleration limitations will greatly reduce the average scanning velocities below the previously stated maximum values. For example, to raster scan a plus or minus two-degree segment sinusoidally, the average scanning velocity in that dimension is reduced to approximately 1.8 degree-per-second, which is only thirty percent of the maximum permissible linear value. The solid angle scanning rate also drops correspondingly to a low of six-tenths degrees squared per second. It should be quite obvious, therefore, that the ability of this particular radar to accomplish volumetric scan is severely limited.

Fortunately, available information indicates that the orbital path of most satellites, and especially those meeting the boundary criteria, can be predicted with a reasonably high degree of accuracy from data which is now supplied by NORAD and the Goddard Space Flight Center. Such data obviously can drastically reduce or negate the requirement for an initial volumetric scan capability for initial acquisition purposes as well as to aid the subsequent tracking function.

If it were possible to predict a satellite's geocentric position as a function of time with an absolute three-dimensional accuracy of four hundred feet or less, all acquisition and tracking in angle could be open loop computer controlled. The resultant angular pointing error would not exceed a maximum of $\pm$ 0.04 degree at slant radar ranges beyond one hundred miles and would improve commensurately as a function of range. It has been indicated that such accuracies may be achievable in the near future using conventional orbit prediction data for satellites meeting the previously described boundary conditions and whose parameters are suitably updated by daily observation. The conclusions of a study conducted by North American Aviation[5] for RADC are even more optimistic. Accuracies of at least two orders of magnitude better than those just mentioned, namely four feet or less, are believed possible if the data is updated from data derived only one or two passes prior to the desired pass. If this conclusion is experimentally verified, not only open-loop angular acquisition and tracking but open-loop range gate acquisition and tracking should also be feasible.

The present predictive accuracy for similarly bounded satellites whose parameters are updated daily is somewhat in doubt, but it is definitely worse than the 400-foot geocentric value previously stated. Maximum cross-trajectory errors of approximately $\pm$ 1/2 mile have been indicated, with in-trajectory errors of from $\pm$ 15 to $\pm$ 30 miles, this last error being commensurate with a three to six second time-of-arrival error. When the parameters have been allowed to "age" for periods up to a week or more without updating (as may well occur in practice), these errors will undoubtedly increase. If one ascribes a linear rate of error increase in both dimensions, an assumption which is not necessarily correct, but appears to be reasonable under the other assumptions made herein, a maximum cross track error of approximately $\pm$ 3-1/2 to $\pm$ 4 miles might be expected after one week, with a time-of-arrival error of up to $\pm$ 1/2 to $\pm$ 1 minute. This latter error would generate corresponding in-track errors of from $\pm$ 150 to $\pm$ 300 miles.

The dimensions of highest anticipated error, which can apparently be orders-of-magnitude larger than the cross track error, lie in the dimension along the path of the trajectory. The otherwise large volume of uncertainty associated with initial acquisition can thus be reduced in practice to a simple two-dimensional problem encompassing the much smaller cross trajectory error if the radar is capable of simultaneously observing all points within the cross-track plane until the satellite passes through.

The antenna beamwidth necessary to ensure simultaneous observation of all points in the cross track error plane is maximum when the slant range is (a) minimum and (b) perpendicular to the cross track error plane, intersecting this plane at the centroid of cross track error. Assuming the latter condition to be the worst look angle situation, the minimum range at which all of the $\pm$ 4 mile cross track error is simultaneously observed is calculated to be 1375 miles. This is somewhat unrealistic for the acquisition of low altitude satellites (i.e., 150-mile altitudes) where the line-of-sight slant range is approximately 900 miles maximum.

The arc intercepted by the 1/3 degree antenna beam at 900 miles slant range is approximately $\pm$ 2.6 miles. Assuming that this is the minimum range at which the satellite becomes visible, the beam must be scanned approximately 0.1 degree to encompass an error of $\pm$ 4 miles in arc. Since the cross track error is two-dimensional, the scan must also be two-dimensional. A circular scan appears to be the most efficient two-dimensional scan to ensure probability of intercept with a scan period of approximately once per second being achievable with the rate and acceleration limitations stated previously. This should practically insure 100 percent intercept under the assumed conditions.

If acquisition must be accomplished at shorter ranges for whatever reason, difficulties may be encountered. A circular or raster scan rate may be too slow to accomplish adequate illumination of the temporally related three-dimensional error space now involved to insure intercept. The most feasible scan method under these conditions, if required, appears to be one which consecutively assumes different cross plane errors as it scans along the trajectory for a sufficient distance to insure intercept within the anticipated, temporally related in-track errors. Such a scan should permit illumination of this potentially large error volume at the highest possible rate. A scan of this type would of necessity require real time computer direction.

Real time computer direction of the antenna beam will also be required in many instances during initial acquisition in order to counteract the adverse effects of the earth's rotation. This rotation continuously changes the topocentric coordinates of the earth located radar site with respect to the geocentrically located satellite orbit. The requirement to continuously point at a particular error volume associated with the anticipated ephemeris thus requires a continuous modification of the radar pointing angle as a function of time. The worst case for the radar occurs when the satellite's anticipated or real acquisition position is due South of the radar position. For a slant range of 900 miles at zero elevation angle, the resultant rate-of-change of azimuth look angle to such a position induced by the earth's rotation would be approximately 0.0164 degrees per second. Although this is inconsequential over a period of a few seconds, 30-second delays are anticipated, for example, which will result in a total pointing error of approximately 0.49 degrees for the condition just described. Such a pointing error can seriously jeopardize the acquisition function, as has been previously pointed out.

A requirement for real time look angle correction therefore exists to insure reliable satellite acquisition by the radar. A computer can most logically accomplish this function; however, its method of implementation greatly influences the type and magnitude of capability required. The tentative approach to be used, but which must be examined in much more detail before its definite adoption, is that which was adopted for this purpose when the radar was originally instrumented for communications satellite acquisition and tracking. This approach permits the utilization of a fairly simple computer to read and interpolate previously calculated look angles in real time for various anticipated early late satellite arrival times along the anticipated geocentrically located ephemeris. This particular operation is far less complex than that required to calculate these anticipated angles, permitting this latter, more complex operation to be accomplished beforehand in non-real time either on the same or a much faster computer.

The present orbital computer is a Packard Bell PB 250 utilizing magnetostrictive delay lines as the basic high access speed memory, and an IBM formatted magnetic tape high capacity memory. Additional peripheral I/O equipment includes a Flexowriter and punched tape reader.

The present instrumentation calculates and utilizes five sets of two-dimensional look angles for every anticipated position of the satellite at one second intervals corresponding to arrival times of zero, ± 30 seconds, and ± 60 seconds. Quadratic interpolation is then used to derive the look angles for one-second temporal error increments. Acquisition is then initiated at the time and angles corresponding to a 60-second late arrival time on the assumption that in almost all instances, the satellite should have arrived earlier. The antenna is then directed to "chase" the satellite by means of the precalculated tracking data. This is accomplished relatively easily by the generation of a dual scanning function which appropriately superimposes on the normal fixed arrival time scanning function the effects of a changing arrival time. The resultant function then causes the antenna to chase the satellite, hopefully acquiring it before the 60-second early arrival function is reached.

One of the chief merits of this approach is that the antenna is moving in the direction of the object's path when acquisition takes place. Two distinct advantages result: (a) the initial antenna acceleration requirements are minimized to establish track lock-on, and (b) the initial monopulse error signal, if used for lock-on purposes, is of the correct polarity to accomplish this function.Otherwise, if the antenna beam were standing still, or if the satellite were "chasing" it, the initial monopulse error signal would be 180 degrees in error. The main disadvantage of this technique, however, is that the time required to complete the acquisition will, in many instances, permit the satellite to approach the radar much closer than the minimum desired 900-mile slant range. The anticipated probability of achieving acquisition would then be expected to drop drastically as previously described.

Fortunately, the same early-late tracking data can be used to permit continuous illumination of a reasonably fixed error volume associated with the satellite trajectory for acquisition as described. This will ensure that the initial slant acquisition range will remain at 900 miles or beyond, as desired, to ensure volumetric intercept with little or no additional scanning required to compensate for the maximum anticipated cross track error. Since the slant radar range to a fixed geocentrically located cross track error position also changes as the result of earth's rotation, some form of range correction function will also be required. The simplest method is to maintain a fixed low elevation look angle during acquisition, effectively limiting the minimum line-of-sight range for a specified target elevation.

During the initial acquisition phase, the antenna beam will be computer-directed toward the particular error volume associated with a one minute early arrival. The elevation angle should also be commensurate with the desired 900-mile minimum slant range or greater. Upon active initiation of the acquisition function, the computer will generate an angle pointing correction function from the pre-calculated data such that the elevation angle remains constant with only the azimuth angle changing to correct for the earth's rotation. The net effect will be to observe a reasonably fixed geocentrically related error volume which slowly slips along the trajectory during acquisition. This slippage results from the combined effects of the fixed elevation angle scan which will be used, the particular azimuth look angle, and the earth's rotation. The rate of slippage under the worst conditions should not exceed approximately 0.2 miles/second, which should definitely be acceptable under the previously calculated scanning conditions.

Monopulse angle error sensing will be used in addition to straight amplitude sensing to determine when tracking should be initiated and what corrections are necessary to establish the desired computer aided tracking function. Various implementation approaches are now being considered.

The first approach is more or less conventional, initially disconnecting the fixed volume computer direction function when the amplitude and error signals are of sufficient strength and initiating interim closed loop monopulse tracking. As previously stated, a polarity sensing technique will be desirable to prevent initial wrong direction acceleration of the antenna and to provide a pre-acceleration capability, if desired. Fortunately, the initial

elevation scanning rates should normally be less than 0.05 degree per second, minimizing the acceleration problem in this dimension. The azimuthal rates, on the other hand, may exceed 0.25 degrees per second. Assuming the previously mentioned wrong direction acceleration can be negated, acceleration to 0.3 degrees per second per second angular velocity can be accomplished in less than 0.1 second from a dead stop. This further assumes a maximum angular acceleration capability of four degrees per second per second which should be achievable. This capability, plus a pre-acceleration capability derived from the aforementioned wrong direction monopulse signal, should generally be more than adequate to permit the desired acquisition function.

Closed loop monopulse tracking would then continue for a sufficient period, i.e., approximately ten to twenty seconds, to permit the generation of a suitable bias to achieve maximum coincidence of the precomputed tracking function with the monopulse tracking function. Open loop computer directed tracking would then be initiated which would be periodically updated and corrected by monopulse signals interspersed on a multiplex basis. Multiplexed operation is a necessity, as described elsewhere, because of the present inability to gather simultaneously high resolution data and monopulse tracking data. Present plans will permit up to one pulse out of every seven to be used for such updating. If the pre-computed tracking functions are sufficiently accurate, however, the sampling period may be extended up to ten or more seconds.

The second approach utilizes the pre-computed tracking function throughout the acquisition tracking process. The satellite is allowed to fly through the near stationary error volume under observation, with monopulse error data being gathered during this period. This position-versus-time data is then used to select the proper pre-calculated two-dimensional tracking function. Once the proper function is automatically determined, which should only take a few seconds, the antenna is directed at a high rate to the calculated, dynamically varying position of the satellite, whereupon it slows to the proper angular tracking velocity and commences the multiplexed high resolution/monopulse error data gathering mode of operation previously described as part of the first approach to correct for errors.

Which of the two techniques will be implemented must yet be determined by further investigation. It is also possible that both approaches may be provided for increased operating flexibility and comparative evaluation.

Regardless of which technique is implemented, a preliminary analysis has indicated the present once-per-second digital readout rate of the present acquisition/tracking computer will not be adequate. From a sampling theory, this rate limits the maximum unambiguous analog bandwidth to approximately one-half hertz. The overall required servo bandwidth, on the other hand, is expected to exceed 10 hertz, if no radome is used. This obvious mismatch can potentially be a great source of error unless suitable correction or compensation is accomplished.

The most straightforward solution would be to appropriately increase the digital computer read-out rate to provide a match to the servo bandwidth. Assuming this bandwidth to be a maximum of five hertz, the minimum digital computer read-out rate must be 10 pulses per second to avoid ambiguous responses. Such a rate is impossible to achieve with the present computational equipment without extensive modification. Although the computational rate of the basic PB 250 computer is more than adequate for this purpose, an additional high speed memory capability is required, or another small computer may be required to take care of the bookkeeping requirements. Both approaches are being investigated, as well as the possibility of complete computer replacement.

Two other somewhat simpler approaches are also being investigated. The first involves a deliberate bandwidth mismatch; for example, a restriction of the computer bandwidth to approximately five hertz. This will naturally generate tracking errors, but preliminary analysis indicates that they may be acceptable. If successful, the required computer read-out rate will be reduced commensurately. The second approach involves the use of a small specialized computer or converter which in essence can take data at the present once-per-second read-out rate and convert it to the desired sampling rate. This would most likely incorporate a power series interpolation function which could then be interrogated at any desired rate. The basic tracking computer would be additionally required to read the various coefficients of the power series into this specialized function at the once-per-second rate. This does not appear to be a problem,

since all necessary computations would be accomplished off line in non-real time. This latter approach is strongly favored at this time as being the simplest and most expedient solution.

The servo bandwidth requirements which have been quoted herein are those which are presently believed necessary to accomplish the acquisition and tracking function with the desired accuracy. Maximum angular accelerations of four degrees per second per second are anticipated during acquisition scan and initial acquisition. When tracking, however, the acceleration is not expected to exceed approximately 0.25 degree per second per second in either dimension with a maximum rate of acceleration change not exceeding 0.03 degree per second cubed. Angular tracking velocities will normally not exceed six degrees per second in either dimension.

The computer-aided angle tracking function will also provide an invaluable memory during possible instances of sporadic signal dropout due to target orientation or ionospheric and/or atmospheric effects. The possible loss of target track should generally be avoided as a consequence. A threshold will be required to reject active monopulse updating data which does not meet the minimum signal-to-noise ratio. It would also command the computer directed tracking function to continue open loop control and perhaps modify the active monopulse updating procedure to temporarily increase its sampling rate until the crisis has passed. A "hold" type of display of the monopulse error signal may also be desirable to monitor the tracking errors, dropouts, etc., and perhaps to permit incremental manual updating if desirable. Additional investigation of these possibilities will be accomplished.

Another important use of the computer-aided tracking function will be the re-acquisition of satellites which cannot be tracked throughout their overhead passage. This tracking deficiency is common to almost all AZ/EL mounts, as compared to equatorial mounts, which cannot plunge through zenith, but must rotate up to 180° azimuthally to continue tracking. The minimum period to rotate 180° azimuthally is a total of 31.5 second, being limited by the aforementioned angular velocities and accelerations. This obviously represents a fairly long period during which the computer-driven tracking function must be sufficiently accurate to permit re-acquisition. Whether this accuracy is achievable remains to be determined, but there is sufficient evidence to indicate that it will be in many instances.

Fortunately, this period of tracking loss resulting from a through-zenith satellite passage represents a worst situation which should occur much less frequently than off-zenith passages which suffer commensurately less tracking dropout. For nominal satellite velocities of five miles per second, for example, there should be no loss of tracking capability, if the normal distance from the radar to the satellite's orbital plane is greater than 50 miles. This corresponds to a maximum elevation angle of approximately 71 degrees during the passage of a satellite at 150 miles local elevation.

The use of computer-aided range and range rate tracking is also contemplated. No modification should be necessary to the present computer program to provide this data except to increase the output data rate beyond once-per-second as previously mentioned. The various techniques which are being considered for angular tracking should also be applicable here. The specialized power series computer is again favored at this time. Active range and range rate will also be derived for direct utilization and/or computer data updating/correction. As previously mentioned in the angular scanning situation, the "coasting" memory capability should be invaluable during those periods when active data is denied for whatever reason.

In conclusion, preliminary investigation indicates that the required satellite acquisition and tracking capability should be feasible as outlined. Additional instrumentation and/or modifications will be necessary, primarily in the computer area, but maximum utilization of existing instrumentation should be possible to minimize such requirements. A more accurate determination of the achievable satellite ephemeris prediction accuracies is extremely desirable, since this factor exerts such an important influence upon the initial acquisition capability. Additional studies are also necessary and will be conducted to firmly determine those modifications and/or new instrumentation which will be required.

APPENDIX C
DISTORTION ANALYSIS OF BROADBAND NETWORKS

One of the major problem areas in the design of wideband systems is the effect on the phase and amplitude of a signal of the dispersive characteristics of components. Among the topics discussed are VSWR, periodic phase and amplitude dispersion, dynamic range, and the properties peculiar to the propagation constant of waveguide. At one point of the discussion the time sidelobes due to two mismatches are shown to be equivalent to a periodic phase and amplitude ripple.

1. MISMATCH EFFECTS

A major factor that determines the optimum performance of the Signal Processing Test Facility is the system time sidelobe requirement. VSWR affects time sidelobe performance as illustrated in the following discussion.

If two VSWRs are represented by reflection coefficients, Γ_1, Γ_2 they will react according to Figure C-1.

It is assumed that the transmission line is lossless. The propagation constant is β, and the mismatches are separated by a distance of ℓ.

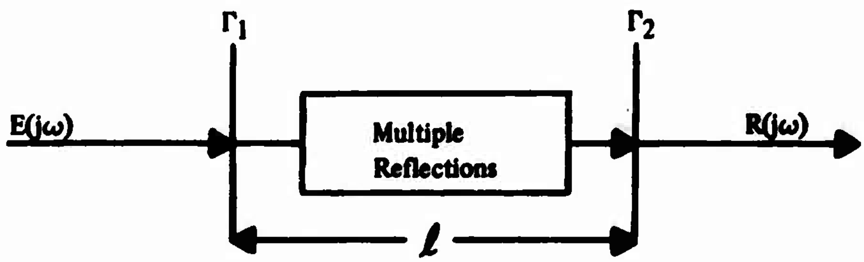


Figure C-1. Spatial Representation of Two Discontinuities

A simple signal flow diagram of this transmission line is given in Figure C-2.

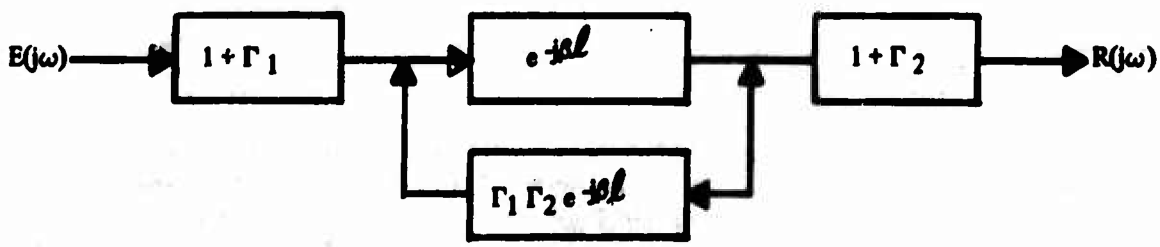


Figure C-2. Flow Diagram Representation of Two Discontinuities

Defining the output prior to any multiple reflection as

$$R_o(j\omega) = (1 + \Gamma_1)(1 + \Gamma_2)\, E(j\omega)\, e^{-j\beta \ell} \tag{1}$$

The transfer function of Equation (1) has the effect of attenuating and phase shifting the input signal by a negligible amount. This can be illustrated by showing that for passive discontinuities

$$[1 - |\epsilon|] < |1 + \Gamma_1 + \Gamma_2 + \Gamma_1 \Gamma_2| < 1 \text{ where } |\epsilon| \ll 1$$

and that the argument of this expression is near zero when the absolute values of the reflection coefficients are small. The exponential in equation (1) has the effect of delaying and dispersing the waveform according to the length and cutoff conditions of the waveguide.

The effects introduced in equation (1) do not appear to be of major concern and so will be neglected in the following analysis. Instead, we will examine the effects of distortion on the signal as produced by the next order of interactions between reflection coefficients.

The first multiple reflection (or echo) is

$$R_1(j\omega) = \Gamma_1 \Gamma_2^{-2j\beta \ell} R_o(j\omega) \tag{2}$$

The output and first echo are

$$R(j\omega) = R_o + R_1 = [1 + \Gamma_1 \Gamma_2^{-2j\beta \ell}]\, R_o(j\omega) \tag{3}$$

The second term on the right hand side of equation (3) may be written as

$$\Gamma_1 \Gamma_2 e^{-2j\beta \ell} R_o(j\omega) = |\Gamma_1 \Gamma_2|\, e^{j\theta(\omega)} R_o(j\omega)\, e^{-2j\frac{\ell}{c}\omega} \tag{4}$$

where the waveguide dispersion has been neglected, i.e.,

$$\beta \cong k = \omega \sqrt{\epsilon \mu} = \frac{\omega}{c},$$

$|\Gamma_1 \Gamma_2|$ is the absolute value of the product of the reflection coefficients, and $\theta(\omega)$ is the angle of that product,

$$\tan^{-1} \frac{Im(\Gamma_1 \Gamma_2)}{Re(\Gamma_1 \Gamma_2)}$$

It can be seen from inspection of equation (4) that the last term is an echo of the original signal bounded above by $|\Gamma_1 \Gamma_2|$ maximum, distorted by the frequency dependence of $|\Gamma_1 \Gamma_2|$ and $\theta(\omega)$, and delayed in time by an amount indicated by the exponent factor $[2\ell/c]$. The phase shift $\theta(\omega)$ has some nominal constant value over the bandwidth of interest. This is not an unreasonable assumption since it can be shown that the phase shift of most types of discontinuities varies only slightly over the useable microwave bandwidth. This is especially true if the discontinuity is mostly susceptive.

In order that the sidelobe level satisfy the 40 db receiver sidelobe requirement the following relations must apply.

$$10 \log \left| \frac{r_1}{r_o} \right|^2 \leq 10 \log \left| \Gamma_1 \Gamma_2 \right|^2 \max < -40$$

or

$$\left| \Gamma_1 \Gamma_2 \right|_{max} < 10^{-4}$$

then

$$\left| \Gamma_1 \Gamma_2 \right|_{max} < 0.01 \qquad (5)$$

Manipulation of $\left| \Gamma_1 \right|$ and $\left| \Gamma_2 \right|$ allows one to establish similar restriction on their reference plane VSWRs

$$S_2 < \frac{(101)\, S_1 - 99}{(99)\, S_1 - 101} \qquad (6)$$

This gives the shaded region in Figure C-3. To use Figure C-3, assume that the VSWR derivable from Γ_1 is S_1 where $S_{max} = S_1 \geq S_2$ and locate its value on the abscissa. Then S_2 may be any value along its vertical intersect within the shaded region. If more than two discontinuities are present, then S_n for $n \geq 2$ can also be found by locating $S_1 = S_{max} \geq S_n$. When $S_1 \leq 1.22$, all S_n being less, the condition for maximum distortion sidelobe is met by the 45° straight line. For values of $S_1 > 1.22$, equation (6) takes over thereby restricting the upper limit on S_n, $n \geq 2$.

For example, if one discontinuity were relaxed so as to permit a VSWR of 1.35, all the other discontinuities in all the subsystems must be tightened to 1.142. It could not be sufficiently predicted how to allocate these different values of VSWR at this time. It was decided therefore to restrict all the VSWRs equally to the nominal value of 1.2.

Waveguide dispersion was neglected in the computation of the above analysis. However, the effect of dispersion will be to further reduce the sidelobe level because the echoes must travel through greater waveguide lengths and are therefore dispersed more than the compensated main lobe.

2. PHASE AND AMPLITUDE DISTORTION

For the purpose of this discussion, we shall define any in-band deviation from a straight line of the phase response as phase distortion. Amplitude distortion shall be defined as the deviation from a line having zero slope of the amplitude response. It is assumed that amplitude weighting will be accomplished in some portion of the system. To make such equalization possible, the effects of the remainder of the components must result in a flat characteristic.

The time sidelobes generated from phase and amplitude distortion over the frequency bandwidth are illustrated below. Since any arbitrary function continuous in its range of definition may be represented as a spectrum of periodic functions, we may discuss the time sidelobe effects of periodic distortions without loss in generality.

Assume a network has some amplitude distortion denoted by

$$A(\omega) = \frac{a_0}{2} + a_1 \cos c\omega \qquad (7)$$

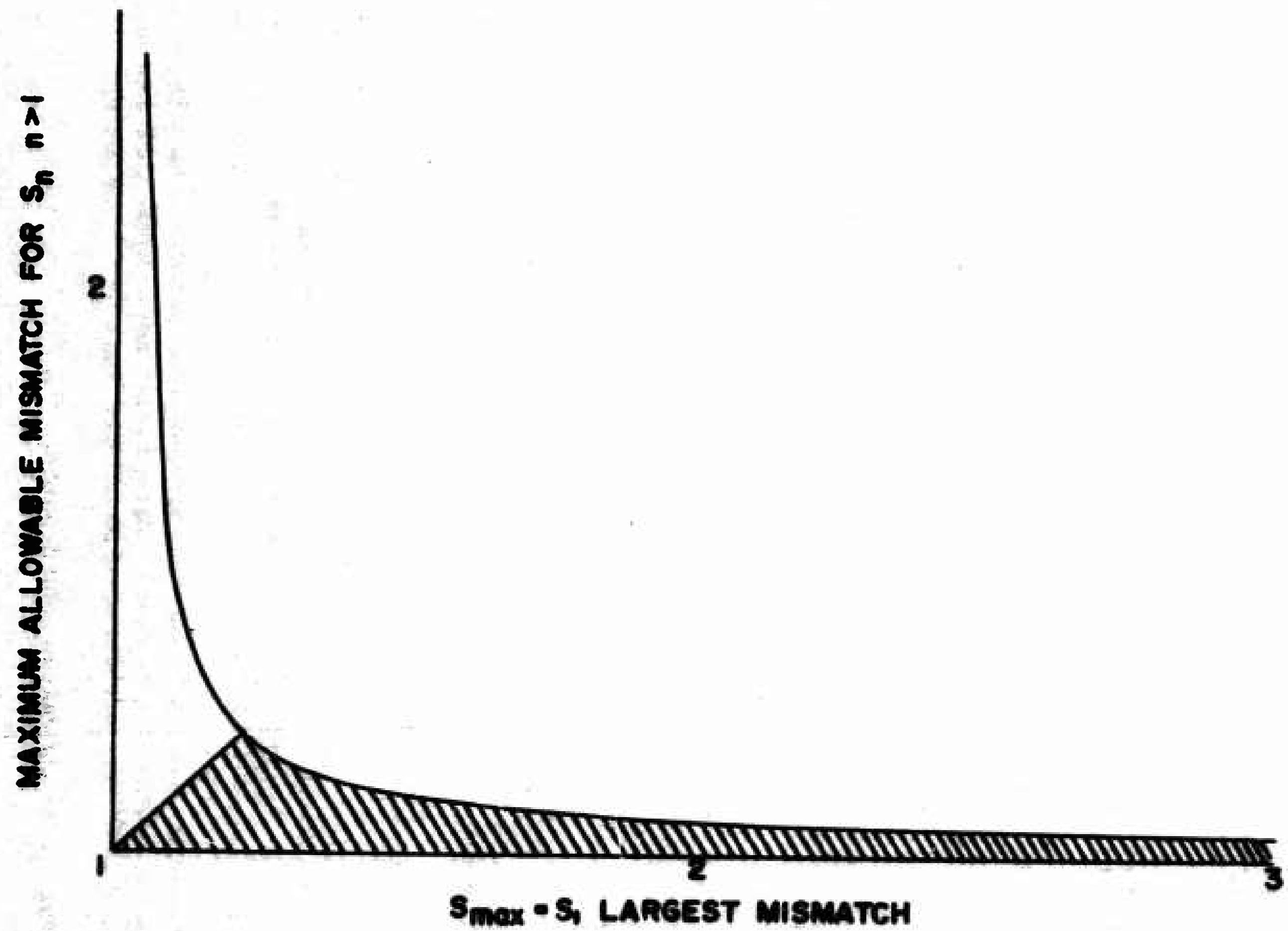


Figure C-3. Upper Limit of VSWR for Each Discontinuity

and some phase distortion denoted by

$$B(\omega) = -b_o\omega + b_1 \sin c_1 \omega \tag{8}$$

The entire transfer function is given by

$$H(\omega) = A(\omega)e^{j\beta(\omega)} H_o(\omega) \tag{9}$$

where $H_o(\omega)$ is the transfer function of an undistorted network.

The impulse response of this function is

$$h(t) = \frac{a_o}{2} J_o(b_1) h_o(t - b_o) + \frac{a_o}{2} J_1(b_1) h_o(t - b_o + c_1)$$

$$- \frac{a_o}{2} J_1(b_1) h_o(t - b_o - c_1) + \frac{a_1}{2} J_o(b_1) h_o(t - b_o + c)$$

$$+ - \frac{a_1}{2} J_o(b_1) h_o(t - b_o - c) + \frac{a_1}{2} J_1(b_1) [h_o(t - b_o + c + c_1) - h_o(t - b_o + c - c_1)$$

$$+ h_o(t - b_o - c + c_1) - h_o(t - b_o - c - c_1)] \tag{10}$$

Frequency responses with the type of distortion represented by equations (7) and (8) describe the behavior of such devices as slow wave structures and all pass networks. The output response of a network described by equations (9) and (10) are shown in Figure C-4.

For small arguments

$$J_o(X) \simeq 1$$

$$J_1(X) \simeq \frac{X}{2} \tag{11}$$

For relatively small values of amplitude and phase distortion, i.e.,

$$0 \leq a_1 \ll 1$$

$$0 \leq b_1 \ll 1 \tag{12}$$

the last term in equation (10) becomes negligible resulting in the following simplification.

$$h(t) = \frac{a_o}{2} \Big\{ h_o(t - b_o) + \frac{b_1}{2} [h_o(t - b_o + c_1) - h_o(t - b_o - c_1)]$$

$$+ \frac{a_1}{a_o} [h_o(t - b_o + c) + h_o(t - b_o - c)] \Big\} \tag{13}$$

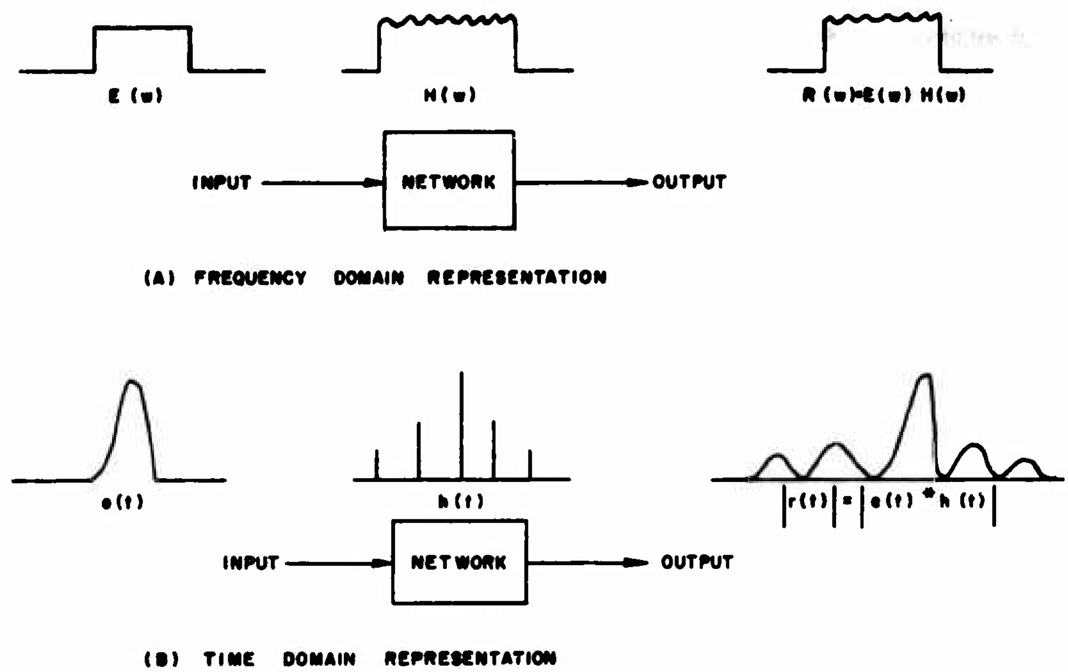


Figure C-4. Signal Representation

The first term in the right hand member is the main lobe shifted by a time b_0 (linear phase term). The second term is a pair of sidelobes of relative amplitude $b_1/2$ and shifted plus c_1 and minus c_1. The third term is a pair of sidelobes of relative amplitude a_1/a_0 shifted plus c and minus c. In order that sidelobes are less than -40 db from the main lobe the following conditions must exist.

$$\left|\frac{a_1}{a_0}\right| < .01$$

$$\left|\frac{b_1}{2}\right| < .01 \qquad (14)$$

If $c_1 = c$ and if $a_1/b_1 < 0$, then for the two pairs of echoes discussed above, the two leading echoes will add out of phase and the two lagging echoes will add in phase, hence

$$\left|\frac{b_1}{2}\right| + \left|\frac{a_1}{a_0}\right| < .01 \qquad (15)$$

The voltages are added in this last expression because the $A(\omega)$ and $B(\omega)$ assumed above result in the same phasor for the time sidelobes. The inequalities are satisfied when the peak phase distortion over the bandwidth is less than .02 radians (1.2 degrees) or when the amplitude dispersion is less than $1/100\ a_0$ (0.1 db).

The last inequality (15) above is of special interest. In order to understand its significance, let us digress to our discussion on mismatch effects. We recall that the first echo was expressed by equation (3) as

$$R(j\omega) = R_o(j\omega)\,[1 + \Gamma_1\,\Gamma_2\, e^{-2jk\ell}]$$

$$= R_o(j\omega)\,[1 + \Gamma_1\,\Gamma_2\, e^{-2j\omega\frac{\ell}{c_o}}] \tag{16}$$

where Γ_1 and Γ_2 are reflection coefficients of each discontinuity, separated by a distance ℓ, and c_o is the velocity of propagation for a distortionless transmission line.

With the same assumption that Γ_1 and Γ_2 are small, the expression in the brackets of equation (16) can be approximated by:

$$[1 + \left|\Gamma_1\,\Gamma_2\right| e^{-2jk\ell}] \simeq [1 + \left|\Gamma_1\,\Gamma_2\right| \cos(2k\ell + \theta)]\, e^{-j\left|\Gamma_1\,\Gamma_2\right| \sin(2k\ell - \theta)} \tag{17}$$

where

$$\theta = \arg \Gamma_1\,\Gamma_2$$

In the discussion on echo produced by two discontinuities we recall that the left hand member of the above expression represents a time sidelobe of amplitude $\left|\Gamma_1\,\Gamma_2\right|$ displaced $2\ell/c$ from the main lobe. Substitution of the parameters of $A(\omega)$ and $B(\omega)$, of equations (7) and (8), respectively, with those of the right hand member of equation (17) gives

$$\frac{a_o}{2} = 1$$

$$a_1 = \left|\Gamma_1\,\Gamma_2\right|$$

$$b_1 = \left|\Gamma_1\,\Gamma_2\right|$$

$$c = c_1 = \frac{2\ell}{c_o} \tag{18}$$

b_o has been omitted since it produces a constant time shift for the entire time response, and is therefore immaterial for our discussion.

The sidelobes produced by VSWRs are in agreement with those produced by the periodic phase and amplitude distortion resulting from such discontinuities as can be seen by the substitutions of the Γ's for the $A(\omega)$ and $B(\omega)$ parameters in equation (15) to obtain (5).

$$\left|\frac{b_1}{2}\right| + \left|\frac{a_1}{a_o}\right| = \left|\Gamma_1 \Gamma_2\right| < 0.01 \tag{19}$$

It should be noted that the limits of periodic phase and amplitude distortion must be jointly specified when their periodicities are identical. This is accomplished more conveniently by establishing maximum VSWRs for each subassembly. Specifying maximum phase and amplitude ripple is applicable for devices that have independent loss or phase shifts not resulting from reflective discontinuities.

This analysis has been carried out only to a first order approximation.

3. DYNAMIC RANGE

If nonlinear characteristics are encountered in any of the receiver components, such as the tunnel diode amplifier, undesired signal degradation may occur. This may take the form of signal suppression, signal distortion and/or spurious signal generation. The range of signal levels over which the aforementioned signal degradation falls within specified limits is commonly referred to as the linear dynamic range of that device or system.

The term "instantaneous" dynamic range is used to describe the capability of a device or system to simultaneously handle two or more spectral components or signals within specified limits of degradation. Manual or automatic gain control may also be used to shift the absolute magnitude of incoming signals to the region corresponding to the optimum "instantaneous" dynamic range. This latter technique serves to increase the overall dynamic range with respect to absolute signal levels, but does not normally affect the achievable "instantaneous" dynamic range.

If nonlinear response of the device or system is expressed in terms of a power-series expansion, the sources of degradation are quite apparent.

$$\begin{aligned} r(t) = g[f(t)] &= C_o + C_1\, f(t) \\ &+ C_2\, [f(t)]^2 + C_3\, [f(t)]^3 \\ &+ \cdot + \cdot \qquad + C_n\, [f(t)]^n \\ &= \sum_{N=o}^{N=\infty} C_N\, [f[t]]^N \end{aligned} \tag{20}$$

The term corresponding to N = 1 expresses the linear gain function. In order to minimize signal degradation due to the other higher order terms, this coefficient should be large compared to the higher order coefficients.

The term corresponding to N = 2 convolves the various signals with one another and with themselves. The first operation results in intermodulation distortion while the second introduces square-law distortion. This effect can be illustrated quite easily using two time-varying signals composed of discrete spectral components. Ignoring absolute phase of any component, since it does not affect the resultant spectral power density, the nonlinear second-order effect upon any one component from each signal can be represented as follows:

$$f_1(t) = a_1 \cos \omega_1 t + a_2 \cos \omega_2 t \tag{21-a}$$

$$[f_1(t)]^2 = \frac{a_1}{2}^2 [\cos (2\omega_1 t) + 1] + \frac{a_2}{2}^2 [\cos (2\omega_2 t) + 1]$$

$$+ a_1 a_2 [\cos (\omega_1 + \omega_2) t + \cos (\omega_1 - \omega_2) t] \tag{21-b}$$

The first two terms represent square law distortion while the last term represents the intermodulation distortion. Although the intermodulation term is undesirable from the standpoint of desired signal degradation in a linear processor such as an amplifier, it is absolutely essential in a linear frequency translator or mixer.

The terms designated by $N \geq 3$ gives rise to cross modulation effects and intermodulation products. (Cross modulation is defined as the modulation of one signal by another.) For example:

$$[f(t)]^3 = [\,- - - + a_i \cos \omega_i t + a_k \cos \omega_k t]^3$$

$$- - - [a_i \cos \omega_i t]^2 a_k \cos \omega_k t$$

$$= \quad [a_k \cos \omega_k t]^2 a_i \cos \omega_i t - - - \tag{22}$$

Since $[a_i \cos \omega_i t]^2 = a_i/2\ [\cos 2\omega_i t + 1]$

The first term expands to

$$[\frac{a_i^2 a_k}{2} (\cos 2\omega_i t) \cos \omega_k t + \frac{a_i^2 a_k}{2} \cos \omega_k t]$$

$$= \frac{1}{4} a_i^2 a_k [\cos (2\omega_i + \omega_k) t + \cos (2\omega_i - \omega_k) t]$$

$$+ \frac{a_i^2 a_k}{2} \cos \omega_k t \tag{23}$$

The first term of this final expansion is an intermodulation effect. If ω_i is close to ω_k, the intermodulation term $(2\omega_i - \omega_k)$ will in many instances fall within the desired signal passband. If a_i and a_k are modulation functions, the resultant product will also modulate the spurious intermodulation signal.

The second term of this expansion is the cross-modulation effect. The term a_i^2 multiplies $a_k \cos \omega_k t$ directly, resulting in a transference of a distorted modulation term (if a_i represents a modulation coefficient) to the other signal.

This same thing will occur for all higher order terms in the Taylor expansion. In order to maintain a low level of cross/intermodulation spurious, it is necessary that the desired or incoming signals be small and that the coefficients c_n be small for all $n \geq 3$ and large for $n = 1$. The coefficient c_2 should be large only if the device is to be used as a mixer.

In order to satisfy the system requirements of $\pm$ 1/4 db amplitude accuracy and $\pm 4^{0}$ phase accuracy, it is necessary that the in-band cross modulation and intermodulation products be maintained below a certain maximum.

For amplitude accuracy:

$$\pm \frac{1}{4} \text{ db} = 10 \log 1.06, 10 \log 0.945 \tag{24}$$

Thus if ΔE is the error voltage due to spurious

$$\frac{E + \Delta E}{E} = \sqrt{1.06} - 1.03 \tag{25}$$

and

$$\Delta E = .03E \tag{26}$$

then

$$20 \log \left(\frac{\Delta E}{E}\right) = -30.45 \text{ db max.} \tag{27}$$

and therefore the total spurious level throughout the system must be below -30.45 db in order to satisfy the $\pm$ 1/4 db amplitude accuracy requirement.

By a similar analysis $\pm 4^{0} = \pm .07$ radians implies an error voltage of $\pm j0.07E$ and

$$20 \log \left(\frac{\Delta E}{E}\right) = 20 \log (.07) = -23.1 \text{ db max.} \tag{28}$$

and therefore the total spurious level must be below -23.1 db in order to satisfy the $\pm 4^{0}$ phase accuracy requirement.

The radar is required to resolve individual target signals ranging over 30 db. The cross modulation products therefore must be less than -60.45 db from the larger signal or -30.45 db from the small signal in order to satisfy the amplitude accuracy requirements. The most extreme condition, the presence of two high amplitude signals, is most apt to upset this -60.45 db requirement. It is believed that the possibility of this occurring is remote and so the specification for cross modulation was relaxed to -20 db maximum from the low level signal, a figure more compatible with the present state of the art. This will result in a maximum accuracy degradation only for large target signals of approximately 0.8 db and a maximum phase inaccuracy of 5.7 degrees.

This relaxation in the requirement implies a 50 db instantaneous dynamic range rather than 60.45 db if the 1/4 db accuracy were to be upheld.

Since the target may vary in range from 100 to 600 miles an additional 30 db dynamic range, not necessarily instantaneous, was added to the specification. This is computed by

$$10 \log \left[\frac{R_{max}}{R_{min}}\right]^{4} = 10 \log [6]^{4} = 31.1 \text{ db} \tag{29}$$

The total dynamic range therefore is 80 db, 50 db of which is instantaneous and the remaining 30 db is adjustable through an AGC or variable attenuator.

4. WAVEGUIDE PHASE DISTORTION

Phase distortion, as defined in section 2 above, due to the waveguide operating over a given bandwidth can be derived from the propagation constant γ which appears in the field equation for the TE_{10} mode.

$$E_y = E_0 \sin\left(\frac{\pi X}{a}\right) e^{-\gamma z} \tag{30.1}$$

where

$$\gamma = \alpha + j\beta$$

The phase constant is

$$\beta = \omega\sqrt{\epsilon\mu}\sqrt{1-\left(\frac{\omega_c}{\omega}\right)^2} \text{ radians per unit length.} \tag{30.2}$$

The distortion of the phase constant, β, about a best fit straight line

$$y = A\omega + B \tag{31.1}$$

is

$$\phi_d = \beta - y \tag{31.2}$$

In order to find the line, y, we form the expectation, E, for the second moment of β about y which may be represented by the notation

$$\overline{\phi_d^{\,2}} = E(\phi_d^{\,2}) = \int_{-\infty}^{\infty} [\phi_d(\omega)]^2\, p(\omega)\, d\omega \tag{32}$$

The function $p(\omega)$ is a density function which in this case we shall define as

$$p(\omega) = \frac{1}{(\omega_2 - \omega_1)} \quad \text{for } \omega_1 < \omega < \omega_2$$

$$p(\omega) = 0 \text{ for all other values of } \omega. \tag{33}$$

The frequency band edges are ω_1 and ω_2 and the density function has been weighted equally through the bandwidth.

The expectation E can now be evaluated

$$E(\phi_d^2) = \int_{\omega_1}^{\omega_2} [\phi_d(\omega)]^2 \frac{d\omega}{\omega_2 - \omega_1} = g(A, B) \tag{34}$$

The expectation (or mean square, as it is sometimes called) will be a minimum when

$$\frac{\partial g(A, B)}{\partial A} = -2E(\omega\beta - A\omega^2 - B\omega) = 0$$

$$\frac{\partial g(A, B)}{\partial B} = -2E(\beta - A\omega - B) = 0 \tag{35}$$

Using the notation

$$\overline{f(X)} = E[f(X)]$$

we may express equations (35) in deterministic form

$$\begin{bmatrix} \overline{\beta} \\ \overline{\omega\beta} \end{bmatrix} = \begin{bmatrix} \overline{\omega^1} & \overline{\omega^0} \\ \overline{\omega^2} & \overline{\omega^1} \end{bmatrix} \begin{bmatrix} A \\ B \end{bmatrix} \tag{36.1}$$

and solving for $\begin{bmatrix} A \\ B \end{bmatrix}$ we have

$$\begin{bmatrix} A \\ B \end{bmatrix} = \frac{1}{|\overline{\omega^n}|} \begin{bmatrix} \overline{\omega^1} & -\overline{\omega^0} \\ -\overline{\omega^2} & \overline{\omega^1} \end{bmatrix} \begin{bmatrix} \overline{\beta} \\ \overline{\omega\beta} \end{bmatrix} \tag{36.2}$$

Assuming that ω is weighted equally within the bandwidth $(\omega_2 - \omega_1)$, we can evaluate the above matrix elements from the integral

$$E[f(\omega)] = \overline{f(\omega)} = \int_{\omega_1}^{\omega_2} f(\omega) \frac{1}{(\omega_2 - \omega_1)} d\omega \tag{37}$$

and we have

$$\overline{\omega^0} = 1 \tag{38.1}$$

$$\overline{\omega^1} = 1/2\,(\omega_2 + \omega_1) \tag{38.2}$$

$$\overline{\omega^2} = 1/3\,(\omega_2^2 + \omega_2\,\omega_1 + \omega_1^2) \tag{38.3}$$

$$\overline{\beta} = \frac{\sqrt{\epsilon\mu}}{\omega_2 - \omega_1}\left[\left(\frac{\omega_2}{2}\sqrt{\omega_2^2 - \omega_c^2} - \frac{\omega_c^2}{2}\cosh^{-1}\frac{\omega_2}{\omega_c}\right)\right.$$

$$\left. - \left(\frac{\omega_1}{2}\sqrt{\omega_1^2 - \omega_c^2} = \frac{\omega_c^2}{2}\cosh^{-1}\frac{\omega_1}{\omega_c}\right)\right] \tag{38.4}$$

$$\overline{\omega\beta} = \frac{\sqrt{\epsilon\mu}}{3(\omega_2 - \omega_1)}\left[(\omega_2^2 - \omega_c^2)^{3/2} - (\omega_1^2 - \omega_c^2)^{3/2}\right] \tag{38.5}$$

These terms are easily computed once the cutoff frequency ω_c is established by waveguide geometry.

For RG-48/U waveguide this value becomes

$$\omega_c = 2\pi\,(2.077852\mathrm{X}10)^9 \text{ radians per second} \tag{39.1}$$

For the frequency range of the radar

$$\omega_1 = 2\pi\,(3.10 \times 10^9), \tag{39.2}$$

$$\omega_2 = 2\pi\,(3.60 \times 10^9), \tag{39.3}$$

$$\sqrt{\epsilon\mu} = 1/c \tag{39.4}$$

where c = velocity of light in feet. Substitution of equations (39) into (38) results in the following values for the matrix elements.

$$\overline{\omega^0} = 1 \tag{40.1}$$

$$\overline{\omega^1} = 21\ 04865 \times 10^9 \tag{40.2}$$

$$\overline{\omega^2} = 443.86834 \times 10^{18} \tag{40.3}$$

$$\overline{\beta} = 16.77144 \tag{40.4}$$

$$\overline{\omega\beta} = 354.08223 \tag{40.5}$$

Substitution of equations (40) into (36.2) solves for A and B, our best fit straight line equation (31.1) becomes

$$y = 8.14155 \times 10^{-9}\, f - 10.503528 \text{ radians per foot of waveguide} \tag{41}$$

The best fit time delay introduced by one foot of RG-48/U waveguide over the bandwidth 3.10 to 3.60 GHz is

$$\frac{dy}{d\omega} = 1.29577 \times 10^{-9} \text{ sec/foot} \tag{42}$$

as compared to

$$\frac{dk}{d\omega} = \frac{d(\omega\sqrt{\epsilon\mu})}{d} = 1.01676 \times 10^{-9} \text{ sec/foot} \tag{43}$$

for free space or a nondispersive transmission line.

This best fit line also has an associated constant phase shift of

$$B = -10.503528 \text{ radians/foot} \tag{44}$$

The distortion of the waveguide

$$\phi_d = \beta - y \tag{45}$$

is determined as

$$\phi_d = 6.388501\sqrt{(f \times 10^{-9})^2 - 4.317470} - (8.14155 \times 10^{-9}\, f - 10.503528) \text{ radians/foot} \tag{46}$$

This has been computed and is shown in Table C-1 for various values of frequency (f).

TABLE C-1

f GHz	ϕ_d (radians/ft)	ϕ_d (degrees/ft)
3.10	-.038211	-2.189
3.15	-.017614	-1.009
3.20	-.002217	-0.127
3.25	+.008391	+0.481
3.30	+.014578	+0.835
3.35	+.016670	+0.9551
3.40	+.014959	+0.857
3.45	+.009708	+0.556
3.50	+.001155	+0.066
3.55	-.010487	-0.601
3.60	-.025022	-1.434

These values are shown plotted in Figure C-5,

Furthermore the distortion is a maximum when f = 3.352 GHz and the phase constant of the waveguide, β, crosses the straight line y at f = 3.20885 GHz and f = 3.5056 GHz. That is

$$\phi_{d_{max}} = \left| \begin{array}{l} \phi_d (2\pi f) = .016673 \text{ radians} = 0.9553 \text{ degrees/ft} \\ f = 3.352 \text{ GHz} \end{array} \right. \qquad (47)$$

$$\phi_d = 0 \text{ at } \left\{ \begin{array}{l} f = 3.2089 \text{ GHz} \\ \\ f = 3.5056 \text{ GHz} \end{array} \right. \qquad (48)$$

It can also be seen that the total excursion in ϕ_d is

$$\left| \Delta\phi_d \right| = \left| \phi_d \max - \phi_d \min \right| = 3.144 \text{ degrees/ft} \qquad (49)$$

across the bandwidth.

If the total amount of waveguide is 65 feet the total excursion in phase over the bandwidth of operation is 204 degrees.

The phase dispersion product is plotted in Figure C-5. The dispersion is mostly quadrat c, and compensation can be provided by simply readjusting the linear FM ramp generated in the signal processor.

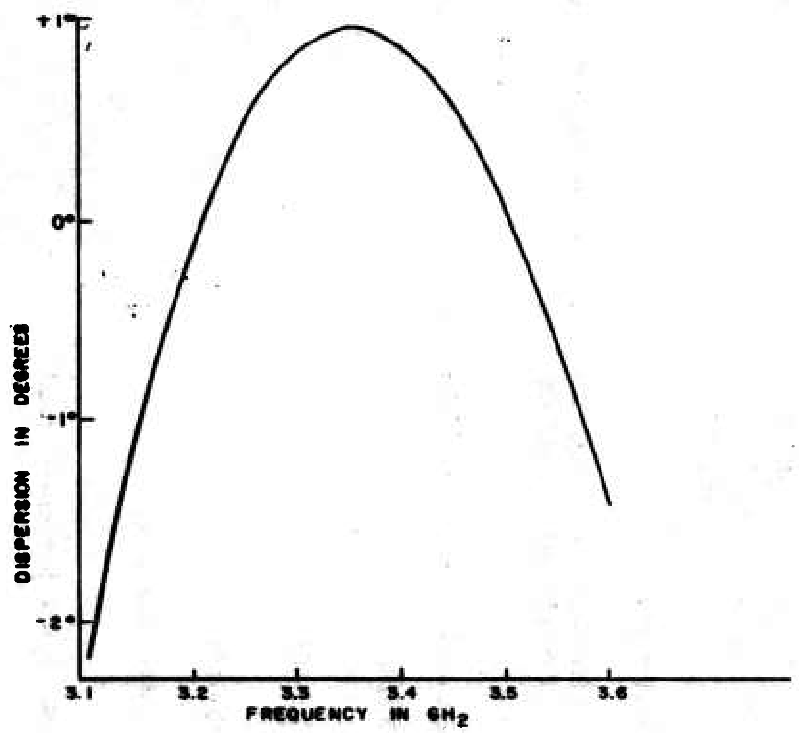


Figure C-5. Phase Dispersion Per Foot of RG-48/U Waveguide

CONCLUSIONS:

The importance of maintaining reflection coefficients below a minimum level has been illustrated. It has been shown that all discontinuities in the microwave transmission line system must be held to a value corresponding to a VSWR of 1.22:1 maximum in order to maintain target echoes below 40 db. If this condition is met any periodic component of the distortion of the amplitude response will be 0.1 db peak and any periodic component of the phase distortion will be 1.2 degrees peak. The above relationships have been verified by paired echo theory.

Since reflection coefficients in transmission line systems give rise to phase and amplitude dispersion transferable to time sidelobes one might consider specifying broadband microwave system performance in time domain parameters rather than the very familiar frequency parameters.

Spacing of discontinuities is also important. Depending on the design philosophy, spacing may be such that no pair of discontinuities ever produces a sidelobe with the same delay as that produced from any other pair of discontinuities. Under these conditions no two sidelobes may add together to produce one of greater magnitude. Spacing of such discontinuities in this manner also results in a pseudo-random phase and amplitude response across the operating bandwidth. Any periodic component of such a response will usually satisfy the conditions for low sidelobe performance if each reflection coefficient is less than 0.1.

Spacing of discontinuities may also be applied to cancel out sidelobes. This can be viewed as a form of equalizing discontinuities with other discontinuities existing within the system.

In cases where distances between discontinuities exceed one half the high resolution range window the restrictions on maximum reflection need not apply.

Isolators and electrically switched phase shifters should be used when necessary to reduce sidelobes.

The discussions carried out in this work are for first order sidelobes only and are valid for reflection coefficients in the order of one tenth. Only sidelobes greater than –40 db were considered.

It has also been shown that sufficient linearity must be maintained for such components as tunnel diode amplifiers and mixers in order that relatively small targets can be recorded within a sufficient level of accuracy.

Finally the total waveguide runs will produce a quadratic phase dispersion having a peak value of 200 degrees maximum and compensation for such must be provided by the signal processing subsystem.

REFERENCES

1. Klauder, J.R., Price, A.C., Darlington, S., and Albersheim, W.J., "The Theory and Design of Chirp Radars" BSTJ Vol XXXIX, No. 4, pp 745-808, July 1960.

2. Wheeler, H.A., "The Interpretation of Amplitude and Phase Distortion in Terms of Paired Echoes", Proc IRE, pp 359-385, June 1939.

3. Goldman, S., "Frequency Analysis, Modulation and Noise", McGraw-Hill, pp. 106.

APPENDIX D
POWER AMPLIFIER

The requirements for the power amplifier for Phase II of the Signal Processing Test Facility place demands on the microwave tubes that significantly exceed the capabilities of existing devices. The operating characteristics of the required amplifier are as follows:

Center Frequency	3350 MHz
Electronic Bandwidth	500 MHz
Duty (RF)	.0028
Pulse Length (RF)	40 microseconds
Gain	87 db
Peak Power Output (mean)	7.15 Mw

Stringent phase linearity requirements are also placed on the amplifier which compound developmental difficulties. The uncorrected value of phase deviation from linearity must be less than 90 degrees over the central 400 MHz of the operating band.

The above requirements are entirely too difficult to be achieved within the one year development schedule established for the program. An interim amplifier will therefore be developed. For this, the requirements for bandwidth, average power, and pulse length will be reduced by a factor of two. The remaining requirements will be essentially unchanged. The principal considerations in the choice of tubes will be as follows:

1. The entire amplifier chain must be operational within a maximum period of six months. This requires that tubes be shelf items, or at the very most, require minor modification.

2. Phase characteristics of the amplifier must be suitable for adequate system performance.

3. The amplifier configuration should be straightforward. Parallel operation of output tubes should be avoided. Space is at a premium and it is desirable that the amplifier be as compact as possible.

4. The type of tubes employed should, if possible, be suitable for phase two operation. This is necessary to preclude major site modifications at the time of changeover.

A product survey was conducted to determine output tubes that could be adapted to Phase I operation. This survey revealed only two suitable tubes. One is a reentrant type, backward wave, crossed field amplifier and the other is a hybrid TWT. The typical characteristics of each of these are listed in Figure D-1 together with the Phase I requirements.

	Crossed-Field Amplifier	Hybrid TWT	Required
Center Frequency	3.0 GHz	2.75 GHz	3.35 GHz
Bandwidth	200 MHz	247 MHz	250 MHz
Peak Power	3.0 Mw	7.2 Mw	7.15 Mw
Average Power	15.0 Kw	10 Kw	14.0 Kw
Pulse Length	10 microseconds	15 microseconds	20 microseconds
Gain	8 db	37 db	———

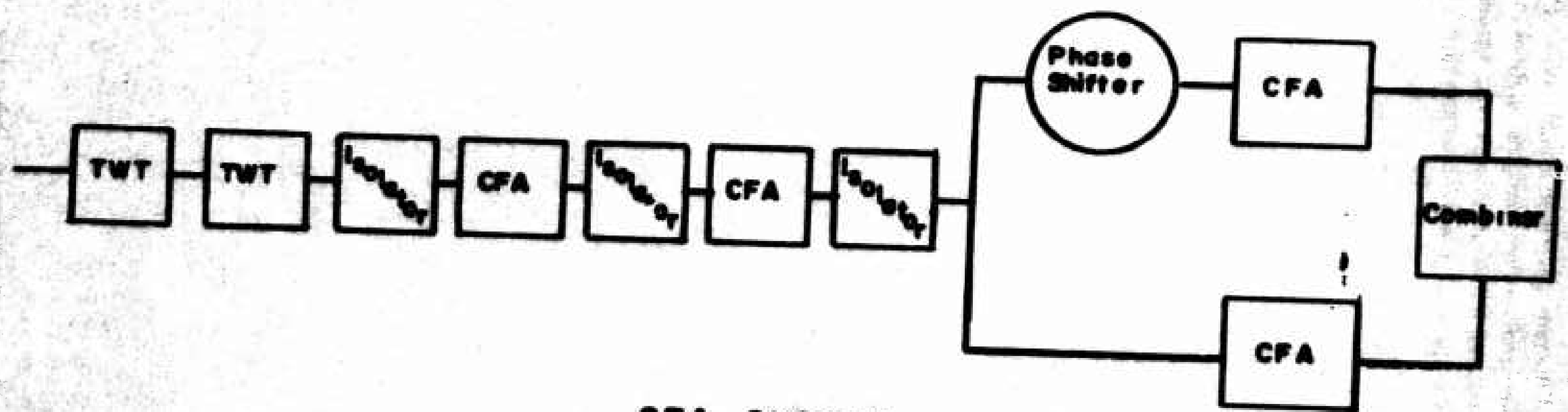
Figure D-1

Due to the fact that the assigned frequency for this system was 3.35 GHz use of either type of tube would require a shift in center frequency and an increase in pulse length capability. These changes could be accomplished within the required six month period. The crossed field amplifier would also require an improvement in electronic bandwidth. This type of change would very likely require considerably more than six months and the confidence that could be placed in developing a suitable tube in the required time would be very low.

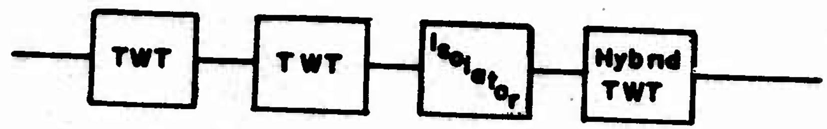
A comparison of the phase characteristics of the two types of tube indicates that the phase sensitivity to voltage change is much superior in the CFA. Published data indicates that a one percent voltage change will cause less than a 0.5 degree change in the CFA and 28.0 degrees in the beam tube. This higher value in the beam tube does not disqualify it, however, since the pulse voltage can be maintained within 0.1%. This results in uncorrected phase ripple of 2.8 degrees which is well within the capability of the system. The phase linearity as a function of frequency is also superior in the CFA, however, the requirement for additional tubes in a series chain would be expected to result in overall characteristics comparable to the beam device. The maximum deviation from linear in the beam tube is expected to be less than 45 degrees. This is near the maximum value that could be tolerated but it is still an acceptable level considering that 30 of the 45 degrees will describe a smooth curve with no more than two complete cycles.

In order to evaluate system complexity, a block diagram employing each type of tube is shown in Figure D-2. The two driver tubes would be the same in either system and would have full Phase II bandwidth capability. A system based on use of the CFA is by far the more complex of the two requiring three additional tubes, a broadband waveguide combiner, and a moderate power hybrid or isolator system. Moreover, the use of parallel output tubes would require tubes with carefully matched characteristics. The CFA system would be expected to be less reliable and more difficult to maintain in operation. It would also require more space which is at a premium on the antenna.

The adaptability of the Phase I tubes to Phase II is, as previously mentioned, a very important consideration. The modification of either type of tube is considered possible although either would represent a major advance in the state of the art. The development of CFA for Phase II would require extensive development work to achieve the required bandwidth and power even if parallel tube operation was employed. If a single tube was to be used, the development would be even more extensive. Likewise, the driver CFA would also require development with the result that several tube developments must be pursued simultaneously. The development of a Phase II hybrid TWT would be more straightforward. The gun, driver, and collector section are within the current state of the art. The major problem would be the output circuit design. The centipede traveling wave circuit has demonstrated an eighteen percent bandwidth at X-band. It is considered capable of the required power although it would require some developmental effort. The circuit can be employed with a high degree of confidence in achieving a suitable tube.



CFA SYSTEM



HYBRID TWT (TWYSTRON) SYSTEM

Figure D-2. Comparison of Amplifier Chains

In summary, the hybrid TWT is recommended for use in Phase I of the radar, based on availability, simplicity of the resulting amplifier chain and adaptability to Phase II modification. Based on the same considerations for complexity and required developmental effort and on the additional consideration that the Phase I system is designed for a linear beam tube, the hybrid TWT (Twystron)[1] employing the centipede structure is recommended for Phase II.

[1] "Recent Advances in the Twystron* Hybrid TWT Art", A.D. LaRue, Varian Associates Publications, Palo Alto Tube Div., Palo Alto, California.

*Trademark, Varian Associates

APPENDIX E
SYSTEM SPECIFICATIONS

The Signal Processing Test Facility will be a high power, wide bandwidth pulse compression radar capable of tracking orbiting space objects and obtaining high resolution amplitude and phase vs range profiles. This information will be recorded on magnetic tape for subsequent analysis. The radar will be completed in two phases to permit experiments at an early date with somewhat reduced performance. The general parameters of the radar are:

	Phase I	Phase II
Peak power	8.5 mw	10 mw
Average power	10 kw	20 kw
Center frequency	3350 MHz	3350 MHz
Modulation bandwidth	250 MHz	500 MHz
PRF	70 pps	70 pps
Pulse width	20 μsec	40 μsec
Unambiguous range window	100 ft	100 ft
Polarization	Transmit vertical Receive - Simultaneous vertical and horizontal in both Phase I and Phase II	

A. FIVE HORN MONOPULSE FEED

The feed assembly design shall consist of a center horn and four peripheral horns arranged to furnish the transmitting and receiving capabilities further described herein.

The center horn shall transmit vertical polarized energy and receive simultaneously both vertically and horizontally polarized signals. During reception, the center horn shall be capable of providing three signal channels. One shall be a narrow band vertically polarized channel for developing the azimuth and elevation tracking reference signal. The other two shall be for the orthogonally polarized 500 MHz bandwidth information signals. The information channels shall be equalized (i.e., balanced) throughout the feed structure so that relative phase and amplitude information of the orthogonal received signals shall be preserved throughout the feed structure.

The peripheral horns which are used only for receiving shall be vertically polarized and shall perform as a narrow band monopulse system with the reference signal being developed in the center horn. The vertically oriented horns shall be used to derive the elevation tracking signals and the horizontally oriented horns to derive the azimuth tracking signals.

(a) Characteristics of Feed Assembly

(1) Center horn

a The center horn shall provide the following:

1 Polarization - Transmit - vertical
Receive - vertical and horizontal

2 Bandwidth - 500 MHz centered at 3350 MHz

3 Power - 10 MW peak and 20 KW average

The transmit channel shall be pressurized. Furthermore, the effects of pressurization on maintaining balance between the two information channels shall be examined. The contractor shall supply a pressurization window for 30 PSIG.

4 VSWR - VSWR measured at the input to the antenna feed system (i.e., input to the transmit channel) shall be 1.1:1 across the operating band.

5 Receive interchannel isolation - Cross coupling between orthogonally polarized signals shall be no greater than 30 db. As a design goal, 40 db coupling is sought.

b The center horn shall be designed such that the broadband information channels shall be connected to RG-48/U hard drawn waveguides which terminate with UG-65/66/U flanges in an existing transmitter-receiver enclosure located on the antenna pedestal. The RG-48/U waveguide runs (approximately 20 feet) and all gaskets, hardware, and support brackets shall be delivered and installed as part of the feed.

1 Both of the entire information channels from the mouth of the horn to the output UG-65/66/U flanges shall be electrically balanced over the operating band such that:

(a) Differential phase dispersion between channels shall not exceed $\pm 3^{\circ}$.

(b) Differential amplitude response between channels shall not exceed $\pm$ 0.1 db.

2 Each channel as defined above shall be designed such that it has the following phase dispersion and amplitude response over the operating band.

(a) Total phase dispersion - less than 260°

(b) Maximum phase ripple - periodic 0.1°
random 2.0°

(c) Maximum amplitude ripple - periodic $\pm$ 0.2 db
random $\pm$ 1.0 db

(d) Insertion loss less than 0.3 db

(2) Peripheral horns

a The four peripheral horns which form narrow band tracking error channels will be used on receive only and shall possess the following operational characteristics.

1 Polarization - vertical

2 Bandwidth - 150 MHz

3 Interhorn isolation - On transmit from the center horn, the energy coupled into the peripheral horns shall be at least 30 db down from the transmit level. The contractor shall consider the effects caused by the energy which is coupled into the narrow band channels. Also the effects of the energy which is not directly coupled but reflected toward the target and received in the information channel as ambiguous target information shall be examined and minimized.

4 Error channel VSWR - The VSWR of the azimuth and elevation error channels looking into the hybrid coupler toward the horns shall be less than 1.1:1 for either channel.

b The outputs of the error channels developed in the hybrid couplers shall terminate in 7/8 inch rigid coaxial line which transmits the signals to the transmitter-receiver enclosure. The contractor shall provide the transmission lines (approximately 20 feet) and installation hardware and install this interconnection as part of the feed assembly.

c The contractor shall terminate the sum outputs of the difference signal combiners with removable dummy loads and provide an additional sum combiner with 7/8 inch rigid coaxial line so that the reference signal from the peripheral horns can be obtained in the transmitter-receiver enclosure. This installation shall be made and tested at the Floyd Test Site then deactivated by removing appropriate jumpers and replacing the dummy loads. Caps shall be provided to protect all unused coaxial fittings from weather or mechanical damage and a suitable storage bracket for wall mounting shall be provided for storage of all unused fittings.

d The entire feed package shall be subjected to outdoor weather conditions as they occur at the Floyd Test Site, Floyd, New York. Consequently, the peripheral horn assembly shall be supplied with weather-proofing radomes.

(b) The feed design shall provide the primary illumination such that the antenna system shall exhibit the following secondary pattern characteristics.

(1) Transmit - The antenna shall have a pencil transmit pattern specified as follows:

a Gain - A minimum of 53 db above isotropic across the operating frequency band.

b Sidelobes - All sidelobes in the radiation pattern shall be at least 20 db down from peak of main beam in both E and H planes.

c Beamwidth - Shall be a nominal value consistent with sidelobe requirements.

(2) Receive

a Broadband information channels shall conform to the following specifications:

1 The orthogonally polarized information channels shall each possess Gain and Sidelobe characteristics with beamwidth considerations as specified in b(1) above. Furthermore, the gain of the orthogonal channels shall provide compatibility with the pre-specified differential amplitude response between channels of $\pm$ 0.1 db.

b Tracking channels - The difference patterns for both tracking planes shall have the following characteristics:

1 Gain - A minimum of 47 db above isotropic.

2 Sidelobes - At least 18 db down from peak of main beam of difference pattern.

3 Null depth - At least 35 db down from peak of main beam of difference pattern.

(3) Boresight accuracy - Using the mechanical boresight of the 60-foot reflector as the reference, and within the operating bandwidth of the channels involved, the electrical boresight shall not deviate more than the following:

a Information channels and tracking reference derived from center horn of the cluster shall not exceed 0.03 degree from boresight reference.

b Tracking patterns derived from peripheral horns of the cluster.

1 Reference pattern shall not exceed 0.05 degree deviation from boresight reference.

2 Difference pattern nulls shall not exceed 0.01 degree deviation from boresight reference.

The characteristics of the Cassegrain reflector configuration in which the feed is to be installed are:

(a) Diameter of primary reflector (parabolic) - 60 feet

(b) Focal length of primary reflector - 25 feet

(c) Diameter of hyperbolic sub-reflector - 69.6 inches

(d) Equation of hyperbolic sub-reflector surface

$$X = 89.4706 \left[\sqrt{1 + (0.0132808y)^2} - 1 \right]$$

(e) Distance of hyperbolic virtual focus from vertex or primary reflector - 66 inches.

(f) Distance of hyperbolic vertex from vertex of parabolic reflector - 272.47 inches.

(g) Half angle at feed position by sub-reflector - 9.2°.

It is noted that a new design for the sub-reflector may be used by the contractor in order to optimize the feed antenna subsystem performance. However, the existing sub-reflector shall be employed unless it can be shown in the proposal that the existing design severely restricts optimum performance.

Dispersion in the feed as well as phase and amplitude balancing in information channels will be given prime consideration. The design of the feed shall be such as to be used in the existing Cassegrain reflector configuration with a minimum amount of modification to existing feed support. Drawings of the feed support and reflector configuration will be provided. The contractor will be responsible for the installation, alignment and boresighting of the feed in the reflector.

B. RECEIVER SUBSYSTEM

The receiver subsystem consists of a four channel receive assembly, high power duplexer, limiters and couplers. Two information channels shall provide a 500 MHz bandwidth capability for the receptionof two orthogonal planes of polarization. The two remaining channels shall perform as narrow band monopulse tracking channels. The broadband 500 MHz channels shall be equalized throughout the receiver path such that relative phase and amplitude information of the orthogonal received signals will be preserved as specified herein. Consideration shall be given to the effects of the duplexer on the relative bandwidth characteristics of the two broadband receive channels. For reference purposes the broadband channel containing the transmit duplexer shall be designated the vertical channel; that not containing the transmit duplexer shall be designated the horizontal channel. All design parameters as specified below shall be compatible with a 40-microsecond system pulse width and PRF of 70 pulses per second.

(a) Characteristics of the Broadband Channels. The 500 MHz information channels shall provide the following:

(1) Bandwidth - 500 MHz at the 3 db points and centered at 3350 MHz.

(2) VSWR - measured at the input of the receive channels shall be 1.20:1 maximum.

(3) Total phase nonlinearity shall not exceed ± 5 degrees. Random phase nonlinearity shall be less than two degrees RMS and the periodic phase nonlinearity shall be less than ± 0.1 degrees peak to peak. These restrictions do not include dispersion due to the propagation constants of waveguide.

(4) Periodic gain ripple shall not exceed ± 0.2 db and random gain ripple shall not exceed 1 db RMS.

(5) Maximum noise figure shall be less than 6.5 db.

(6) Channel gain shall be at least 18 db.

(7) The overall receiver/signal processing configuration shall exhibit sufficient dynamic range such that undesired signal suppression and/or the generation of spurious signals shall not occur. The dynamic range shall be sufficient to handle individual target signals whose difference in energy may exceed 30 db, and whose absolute energies which are generated shall be at least 20 db below the minimum target signal after pulse compression in order to permit attainment of a ± 1/4 db amplitude accuracy and ± 4 degrees phase accuracy.

Thus, the receiver chain shall exhibit an instantaneous dynamic range (in the presence of two or more maximum energy signals to that range) of at least 50 db. The maximum expected uncompressed signal power level is -60 dbw for a 1000 square meter target at 100 miles range. It is expected to diminish by at least 30 db as the range increases up to 600 miles.

The overall dynamic range, therefore, in the receiver chain shall be at least 80 db, however, only 50 db of this value shall be of an instantaneous capability. The additional 30 db of dynamic range can therefore be achieved by a suitable gain control derived from the range and/or signal function.

The tunnel diode shall be capable of handling the maximum expected signal level. However, it may not exhibit the required 50 db dynamic range simultaneously over the expected -60 and -90 dbw maximum expected signal level. Thus, suitable attenuators may be required preceding the tunnel diode amplifier.

(8) Gain stability $\pm$ 0.5 db/hr maximum.

(9) Phase stability $\pm$ 1°/hr maximum.

(10) Minimim power handling capability in the receiving path shall be 20 kw peak and 40 watts average for fail safe operation.

(11) Recovery time from the transmitter mode to the receive mode shall not exceed 30 microseconds.

(12) Differential phase dispersion between channels shall be less than $\pm$ 1° for any frequency within the bandwidth with the same input and output loads. In the case of the random component of differential dispersion an RMS value not exceeding one degree shall apply.

(13) Differential amplitude response between channels shall not exceed $\pm$ 0.3 db with $\pm$ 0.1 db as a design goal for any frequency within the bandwidth with the same input and output loads.

(b) In addition to the above, the vertical channel shall provide the following transmit capabilities:

(1) Transmit Power - 10 Mw peak, 20 kw average.

(2) Transmit - Receive duplexer operation with a recovery time not to exceed 30 microseconds.

(3) Insertion loss from the transmitter to antenna ports shall be less than 0.5 db.

(4) Spike leakage and isolation from the transmitter to receiver ports shall be reduced to a level which shall result in no permanent degradation in receiver performance. Receiver protection shall be provided against spurious effects such as voltage breakdown to assure that reflection of transmitter power does not damage the equipment.

(5) Pressurization and cooling shall be sufficient for obtaining the specified power handling capability.

(6) VSWR of the transmitter port shall be 1.15:1 maximum.

(7) Reflected transmit power indication. Rapid increases in reflected power as would result from breakdown of the transmit signal shall be indicated by low voltage video signals. These signals shall indicate the reflected transmit power at the output of the transmitter path and the transmit power directed through the duplexer and into the receiver.

(c) The broadband receive channels shall each be provided with a signal monitoring and driving capability of 40 db $\pm$ .5 db coupling across the bandwidth. In the case of the vertical channel the monitoring and driving capability shall be provided for the transmit as well as the receive path. 40 db bi-directional couplers with 20 db directivity located at the antenna ports of these channels are recommended.

The broadband receive channels shall terminate with an amplifier limiter device preferably a tunnel diode amplifier preceded by a limiter. There will be no frequency conversion in the broadband channels. The broadband information channels shall incorporate UG-65/66/U type flanges for the input ports. These flanges are designed for RG-48/U waveguide. The output connection will be specified at a later date.

(d) Noise Figure Monitor - Automatic noise figure measurement capability shall be provided across the receive channels terminating at the IF frequencies. One set of Noise Figure Measurement equipment selectable for any one of the receive channels will be necessary.

The noise figure monitor shall be capable of readout of the noise figure of either of the four receiver channels. The readout shall be directly in db and shall not affect the performance of the radar receivers. ' ;as tube noise source coupled to the channels shall be gated on during a portion of the pulse repetition period and provide a calibrated reference noise level. The calibrated noise level shall be compared to the receiver noise level by automatic circuitry the output of which shall be calibrated directly in receiver noise figure.

(e) The narrowband monopulse error channels shall provide the following capability. Each of the two channels shall be identical. They shall each terminate in an IF frequency.

(f) The RF portion of each narrow band channel shall have the following capability:

(1) Bandwidth shall be 150 MHz at the 3 db points and centered at 3350 MHz.

(2) VSWR at input 1.15:1.0 maximum.

(3) Gain shall be 18 db minimum.

(g) The narrow band monopulse receiving channels shall provide the following overall capabilities:

(1) Noise figure shall be 7.5 db maximum.

(2) Gain shall be 12 db minimum.

(3) Dynamic range shall be 45 db minimum.

(4) Gain stability - ± 0.5 db/hr.

(5) Phase stability - ± 1°/hr.

(6) These channels shall withstand a 10 kilowatt peak and a 20 watt average RF power input without causing permanent damage to the low signal sensing devices.

(7) Recovery time from the high power to the low sensitivity state shall be one microsecond maximum.

(8) Output frequency shall be 70 MHz (L.O. Signals will be supplied as required.)

(h) The narrow band channels may terminate with standard type mixers for conversion to the IF frequency. The same conditions for monitoring and driving the receive signal shall apply in these channels; 40 db ± 0.5 db coupling coefficient and 20 db minimum directivity. These channels shall incorporate a 7/8" coaxial flange at the input ports.

(i) Power supplies necessary for operation of the receiver subsystem diode amplifiers, limiters shall be provided by the contractor. The site will provide three phase 208 volts, 60 cycles prime power at the receiver location.

(j) Adequate shielding of the receiver components shall be provided to insure that no degradation in receiver performance shall result from operation within the same enclosure with the transmitter.

(k) The operational temperature specification required is 72°F ± 5° and in the nonoperable condition is -30°F to +100°F.

C. WIDEBAND SIGNAL PROCESSOR

The wide bandwidth portions of the signal processor will be co-located with the driver and final amplifier of the transmitter in the enclosure directly behind the antenna feed structure in order to minimize the length of the RF transmission line.

The signal processor will perform several functions in the final radar configuration.

(a) Provide waveforms to the transmitter driver for several modes of operation and coherent reference signals to the receivers.

(b) Receive and transform the wideband signals to a narrower bandwidth for subsequent recording and analysis.

(c) Provide synchronizing and timing signals to the remaining subsystems to insure proper sequencing of the overall radar system.

(d) Compensate for phase and amplitude distortions in the complete system to insure adequate range sidelobe performance.

The portions of the signal processing subsystem not directly requiring the handling of the total signal bandwidth will be located in either one of the other enclosures on the back of the antenna structure or within the operations building. All essential operating functions will be remoted to the operations building located approximately 100 feet from the antenna pedestal.

The Signal Processor Subsystem (SPSS) shall be capable of generating the waveforms to be transmitted, processing all received signals including angle tracking signals and providing all necessary coherent reference signals and synchronizing pulses to be used by the complete radar system. The SPSS shall be capable of operating in the following modes with provision for switching between them as required.

Mode I: The transmitted waveform shall be a cw pulse at 3350 MHz ± 50 MHz center frequency and 20 μs duration. This signal is to be used for search, acquisition and monopulse angle tracking. This mode shall be tunable over 100 MHz centered at 3350 MHz.

Mode II: The transmitted waveform shall be a linear FM pulse at 3350 MHz ± 100 MHz center frequency, 2.5 MHz modulation bandwidth and 20 μs duration. This signal will be used for intermediate range resolution and range rate determination. This mode shall be tunable over 200 MHz centered at 3350 MHz.

Mode III: The transmitted waveform shall be a linear FM pulse at 3350 MHz center frequency, 250 MHz modulation bandwidth and 20 μs duration. This signal will be used to obtain the high resolution amplitude and phase vs range profiles which are the required outputs of the signal processor.

Mode IV: The transmitted waveform shall be a selectable sequence of signals from Modes I, II, III, or any combination of these modes. The sequence to be used shall be completely flexible as to choice of individual waveforms and programming of the sequence.

(a) Signal Generation: The signals specified above shall be generated and supplied to the transmitter driver with associated synchronizing pulses at a peak power level of 20 milliwatts and impedance of 50 ohms.

(b) Signal Reception and Processing: The SPSS shall be compatible with the other associated radar components as specified herein. The RF outputs from the tunnel diode preamplifiers shall be accepted with subsequent frequency translation and processing provided to obtain the performance required herein. The input signal-to-noise ratio shall not be degraded. The general processing required in the various modes is as follows:

Mode I: The received pulse shall be processed to result in signals suitable for target acquisition and utilization of the azimuth and elevation difference signals for angle tracking. The signals to be received and processed in Mode I will be derived only from the transmitted polarization.

Mode II: The signals shall be processed and compressed to result in signals which can be utilized for intermediate range tracking and range rate determination. The Mode II signals will be derived only from the transmitted polarization. AGC shall be provided to cover the required dynamic range of approximately 85 db.

Mode III: The high resolution signals shall be received and processed from both the transmit polarization channel and the orthogonal polarization channel. The high resolution processing shall result in amplitude and relative phase versus range (within the window) profiles for both polarization at an output bandwidth not exceeding 2.5 MHz. It is further required that all four profiles shall be available at a single output channel. A maximum utilization of common components in the two high resolution channels shall be accomplished.

Mode IV: Appropriate combination of Mode I, Mode II, and Mode III processing.

(c) Coherent References: All reference signals required for frequency translation and coherent detection shall be derived within the SPSS from the station clock and frequency standard.

(d) Range Window: The implementation of the SPSS shall result in a minimum unambiguous range window of 100 feet.

(e) Window Distortion: When operating in Mode III, the distortion across the range window of the processed signals shall not exceed the following:

(1) Amplitude: Plus or minus 0.5 db.

(2) Phase: Plus or minus 1.0 degrees.

This distortion is inclusive of that resulting from the use of weighting filters.

(f) The maximum range sidelobes from a point target for:

Mode II - Shall be at least 30 db below the main response.

Mode III - Shall be at least 35 db below the main response for the duration of the range window.

(g) Range Resolution: The Mode III operation of the SPSS shall provide resolution of four feet or less for two equal velocity point targets differing in amplitude by 34 db and located anywhere within the range window.

(h) Doppler Compensation: Doppler or range rate measurement and compensation shall be provided to insure adequate range accuracy and sensitivity to perform as required in all four modes. The radial velocity component shall be compensated in Mode III to an accuracy of plus or minus 200 feet per second. Initial velocity information will be provided from space track data.

(i) Doppler Readout: Provision shall be made for visual and digital readout of radial target velocity on a continuous basis to be remoted to the operating console.

(j) Data Readout: During Mode I and Mode II operation, continuous output of video signals is required with accompanying synchronizing signals for visual display on conventional oscilloscopes at the operating console. In Mode III, amplitude versus range and relative phase versus range for both horizontal and vertical polarization channels shall be provided from a single output connector in sequential fashion with appropriate synchronizing and identification signals. The output from the signal processor shall be five volts peak across a 50 ohm impedance.

(k) Timing Signals: The station clock will supply the SPSS a system trigger of 100 volts across 50 ohms for a duration of 2 usec. This basic system trigger will be at a PRF of 70 pps. All other timing and synchronizing signals required by the SPSS shall be derived internally from the system trigger.

(l) Distortion Compensation: Capability shall be provided by the SPSS for pre-distortion of the Mode III transmit signal to compensate for 60 feet of waveguide (RG-48/4). Capability shall also be furnished for compensation of fixed repeatable phase distortion within the total system not to exceed 20 degrees peak. Further provision shall be made for measuring and actively compensating for slowly varying phase distortion (at the PRF rate or lower) not to exceed 10 degrees peak.

(m) Dynamic Range: The SPSS shall be designed to have a usable dynamic range corresponding to the front end tunnel diode amplifier dynamic range of 45 db and the additional dynamic range due to processing gain and any changes in gain levels that may be used in attaining this dynamic range during operations shall be recordable. The gain level shall be constant over the range window. The instantaneous dynamic range of the SPSS shall be 45 db minimum.

(n) Channel Matching: The high resolution channels for vertical and horizontal polarization shall be matched to within $\pm$ 0.5 db in amplitude and $\pm$ 0.5 degrees in phase.

(o) System Stabilities: The overall SPSS time and frequency stabilities shall be such that the performance requirements are met even though the tolerances are not specifically noted herein.

(p) Switching: Provision shall be made for all necessary switching controls to change operating modes both from the equipment location and the control console in the operations building.

(q) Remoting: Instrumentation shall be provided for complete operation and monitoring of essential functions of the SPSS from the control console in the operations building.

(r) System Checkout: The SPSS shall have a closed loop capability for testing and adjustment. Low power checkout at the contractors plant shall include a 35-foot waveguide run from the waveform output to the signal processor receiver input to demonstrate capability to compensate for degradation caused by microwave components. A bidirectional coupler is being provided for transmit signal sampling on site.

(s) Angle tracking receivers shall be provided which are compatible with the five horn monopulse feed. The receivers shall develop azimuth and elevation error signals from the elevation and azimuth difference channels in conjunction with a reference (sum) channel. The difference channel inputs will be provided (under separate procurement) at IF (approximately 70 MHz). The vertically polarized sum channel shall be derived within the SPSS. The SPSS shall also provide coherent and isolated local oscillator signals for the difference channel mixers. The error signals shall be supplied to the existing servo cabinet at 30 volts peak-to-peak across 50k ohms with a sensitivity of three volts per degree.

(t) Phase II Modification: The design, development and fabrication of the Phase I SPSS shall be such that a minimum of component replacement and modification will be required for the subsequent transition to Phase II operation.

(u) Special Test Equipment: Any special tools or test equipment required for operation and maintenance of the SPSS shall be supplied as part of said equipment.

RANGE TRACKER:

(a) Analysis to determine the design criteria of a range tracker to be incorporated into the SPSS.

(1) The goals of the range tracker are to track with accuracies such that the radial motion of the gross target is removed, i.e., the range window jitter about the target center of mass is effectively minimized and is a fraction of the radar resolution cell.

(2) Consideration shall be given to the following individual or combined modes of range tracking:

a Pre-computed programmed track

b Automatic track

c Programmed track with real time upgrading

d Doppler aided range track

(b) Design and fabrication of the range tracker in accordance with the following:

(1) The range tracker shall be capable of automatic track on an orbiting space object with a signal-to-noise of 3 db at the maximum expected range rate.

(2) Memory range track shall be provided when the signal-to-noise falls below automatic tracking sensitivity.

(3) Range tracking capability shall be furnished to result in an orderly and accurate transition between various operating modes.

(4) The maximum jitter of the overall system in Mode III shall not exceed ± 0.10 ft. exclusive of signal-to-noise considerations.

Phase II modifications shall be designed to upgrade the performance of the SPSS to the following:

(a) Pulse width, (all modes) 40 usec.

(b) Mode III bandwidth - 500 MHz linear FM.

(c) Mode III resolution two feet or less for two equal velocity point target differing in amplitude by 34 db and located anywhere within the range window.

(d) Compensation of transmit or receive signals for atmospheric dispersion as a function of antenna elevation angle.

D. WIDEBAND TRANSMITTER

The wideband transmitter shall be designed to be located behind the 60 foot parabolic antenna presently located at the RADC Floyd Test Site.

The enclosure to be used in this modification shall be environmentally controlled. In the layout of positions for the various components of the transmitter, space shall be reserved for those components having a direct bearing on the phase stability of the RF signal. Space shall also be reserved for portions of the receive system which will require a controlled environment. An additional enclosure is available for components (such as the modulator) which has no direct effect on the phase and amplitude stability of the system.

The transmitter shall be initially implemented with the type VA145(F) Twystron to operate in accordance with the initial performance parameters. The transmitter shall be capable of being converted, with a minimum of modification, to operate the full power Phase II tube in accordance with the final performance parameters.

Performance Parameter	Initial	Final
Frequency	3350 MHz	3350 MHz
Bandwidth (2 db)	250 MHz	-
Minimum Instantaneous Electronic Operating Beamwidth	250 MHz	500 MHz
Band Edge Power (500 MHz Bandwidth)	-	-2 to -6 db
Band Center Power (Center 325 MHz)	-	0 to -2 db
Band Center Power Variation from Smooth Curve	-	± 1/2 db
Peak Power Output (Max)	4.5 to 8.5 MW	10 MW
Min. Mean Peak Power Output (Linear F.M. Ramp)	5.5 MW	7.5 MW
Pulse Repetition Rate	70 pps	70 pps
Pulse Width (R.F.)	20 microseconds	40 microseconds
Beam Voltage (Max)	115 KV	180 KV
Beam Current (Max)	131 A	152 A
Beam Current (Min)	116 A	133 A

The contractor shall be responsible for demonstrating and proving the performance parameters of the initial designs using a VA145F Twystron. A clear demonstration that the equipment shall be capable of accepting the development model, Phase II tube when it becomes available, and be able to meet the requirements is outlined under the final performance parameter and the phase frequency transfer characteristics to be outlined.

The transmitter shall include the following items:

(a) Power Amplifier

(1) Oil insulated cathode tank sealed against moisture and leaks but with thermal expansion facility to contain cathode bushing, filament transformer, and pulse transformer for Twystron.

(2) Magnetic focusing solenoid for Twystron. To be mounted on cathode tank.

(3) VA145 Twystron amplifier tube to fit within the preceding two items. The VA145 Twystron Amplifier tube shall be supplied with a DC filament and well filtered DC supply with a minimum reserve rating of 30% over the VA145 requirements.

NOTE: If this item cannot be operated in any position from vertical over into a horizontal position, then this and the preceding two items (cathode tank and solenoid) shall be mounted in gimbals to enable the tube to remain in a vertical position while the enclosure in which these components are mounted tilts with the antenna dish.

(4) Fast acting protective equipment to insure that arcs occurring within the tubes or in the output waveguide do not result in destruction energy release to the detriment of the Twystron. Arcs on five consecutive pulses shall result in transmitter shutdown. Other arcs less frequent shall only initiate action to divert remaining energy in the pulse away from power amplifier tube. Fault sensing elements to be used will include current transformers to monitor tube arcs and waveguide arc detectors, both optical and reflected power sensing couplers. These sensors shall be rated so that saturation does not occur and prevent positive operation. Protective equipment will accept fault signals from at least three (3) individual waveguide fault sensors.

(5) Waveguide sections and transitions as needed for connection of a power amplifier tube to duplexer and line to antenna feed. H-Plane bends shall not be used due to resultant excessive dispersion. If conditions mentioned in Note, paragraph Da(3) prevail, appropriate rotary joint shall be provided.

(6) X-Radiation. All electronic or electronic devices capable of producing X-radiation will be so designed, fabricated, shielded and operated so as to avoid overexposure of personnel. For the purposes of equipment and installations design, shielding requirement shall be maintained which limit the X-radiation to less than 2.5 milliroentgens per hour (mr/hr) within one foot (12 inches) of any accessible area, of the radiating source, normally inhabited by personnel during installation, calibration, maintenance, repair and operation of the equipment. The 2.5 mr/hr level is based on a maximum permissible exposure of 100 mr/week. X-ray shielding enclosure for the Twystron must be adequate for the full power development tube.

(7) Handling fixtures shall be provided to enable precise position control and alignment of power amplifier tube during replacement in solenoid with external assistance only of the presently assigned Hi-Ranger Model 10.88 "cherry picker". The time for tube replacement should not require in excess of two hours with the services of three skilled technicians.

(8) Radio frequency dummy load capable of dissipating the full output of the amplifier (both peak and average) to enable transmitter power measurements by calorimetric methods for calibration of couplers. Means shall be provided to enable switching the load in and out of the circuit without disassembling any transmission line installations. The inclusion of this feature shall not noticeably degrade the VSWR of the amplifer operating into the duplexer antenna line. The VSWR of the dummy load shall not exceed 1.15 to 1 over the range of 3100 to 3600 MHz.

(9) All microwave components other than the final amplifier tube shall be designed to meet the final performance parameters.

(10) A VSWR at the output window of the power amplifier in excess of 1.4:1 shall initiate the removal of RF drive to the Twystron.

(11) Harmonic radiation from the transmitter will be reduced to a minimum of 20 db below the fundamental frequency at any point across the operating bandwidth.

(b) High Power Modulator

The modulator shall be located in the enclosures on the rear of the antenna. The components selected must be capable of operating in any position from standing up (to vertical) to laying on their sides (horizontal).

(1) The modulator shall operate with the following utilities already in place on the site.

a Supply voltage (from existing power supply) 0 to 20 KV direct current (continuously variable).

b Maximum available current - 60 amps.

c Three-phase 60 cycle power at 208 volts four wire available in the enclosures for powering auxiliaries.

d There is adequate circulation of liquid coolant in the enclosures at the back of the dish.

e Synchronizing pulse of 50 volts on a 50 ohm line. The synchronizing pulse shall be timed to center the RF drive pulse within the modulator pulse of 25 or 45 usec.

(2) No variation in pulse repetition rate is contemplated or desired. It is pointed out that the change in operation from the initial conditions (25 microseconds) to the final condition (45 microseconds) shall be a one time change accompanied with the replacement of the power amplifier tube.

(3) The modulator shall be designed to enable sustained operation for 23 hours per day on an indefinite basis.

(4) The modulator shall include at least:

a DC resonant charging choke and holdoff diode.

b Pulse forming network built for accommodation of both 25 and 45 microsecond pulse widths. Separate 25 and 45 microsecond pulse forming networks, for Initial and Final operation respectively, shall be acceptable if required to achieve optimum performance.

c Thyratron

d Pulse Transformers

e Backswing clipper diode

f High voltage crowbar circuit shall be incorporated in modulator to divert energy in the cathode pulse away from the Twystron cathode in the event of tube or system arcs.

g Pulse monitor circuit to enable remote viewing of modulator charging waveforms, output pulse waveform to power amplifier cathode.

h Self adjusting pulse leveler circuit to remove ripple from top of output pulse to a maximum of 0.03 percent as an objective. The phase requirements of Dc(3) shall take precedence over this requirement.

i Video load for modulator to enable full power modulator testing without the power amplifier tube. This mode of operation may be restricted to vertical position only (antenna beam elevation at horizon) if desired.

j R.F. filtering/shielding shall be provided so that transmitter spurious r.f. radiation will not interfere with or limit radar operation.

k An inverse clipper diode across the PFN to limit negative voltage on the PFN due to load mismatch.

(c) Driver

(1) The driver shall be located within the existing air conditioned enclosure on the rear of the antenna dish.

(2) It shall have the capability to provide sufficient RF drive to the VA145 or the VA915A. This drive level may be amplitude shaped in order that the phase frequency and the frequency power output amplitude characteristics of the Twystron may be used to help give the specified performance specifications of the transmitter across the 500 MHz operating bandwidth. The excitation input drive level to the driver stage is 20 milliwatts on a 50 ohm line.

(3) Let the phase error as a function of frequency be defined as the difference between the actual phase and the phase described by a corresponding point on a straight line, where that straight line is defined as the locus which makes the integral of the phase error (integrated over the total band) equal to zero. Then, providing that the developmental model power amplifier tube, VA915A, meets the following phase performance characteristics:

Phase Performance Characteristics of the VA915A.

The phase versus frequency characteristic may be considered to be comprised of two distinct variations. A smooth "base curve" and a higher frequency deviation from this curve. The smooth "base curve" will be continuous across the 500 MHz of operating bandwidth with no reversals in slope of the phase (ϕ) versus frequency (f) characteristic describing the curve, and with no more than two complete cycles of deviation from a linear curve of phase versus frequency. (Reversal in slope is defined by the condition $d\phi/df < 0$, i.e,, negative.)

The maximum phase deviation of this smooth base line from the above defined straight line shall not exceed 30° over the middle 400 MHz of the frequency band. Over the remaining 50 MHz at each end, the phase deviation shall not exceed 50°. The higher frequency variation shall not exceed more than 20 complete cycles over the entire 500 MHz bandwidth. The maximum phase deviation of this variation shall not exceed more than 15° from the smooth base curve defined herein.

The phase transfer function of the entire transmitter from driver input to power amplifier output shall be as follows:

a Single pulse phase error

1 The phase error within the middle 400 MHz of the band shall not exceed $\pm 12^{\circ}$ and have an objective of $\pm 10^{\circ}$.

2 The phase error within the upper and lower 10 percent of the operating band shall not exceed $\pm 30^{\circ}$.

3 The average phase error (when averaged over any 5 MHz interval within the middle 400 MHz of the band) shall not exceed $\pm 6^{\circ}$ and have an objective of $\pm$ 5 degrees.

b Pulse-to-pulse phase error

After suitable warmup period has elapsed, the average phase error (as defined in Item Dc(3) a 1 above) of a group of pulses shall not differ by more than $\pm 1^{\circ}$ from the average phase error of any other group of pulses occurring during a 150 millisecond period with a two (2) hour interval between the 150 millisecond sampling periods.

If the phase performance characteristic of the VA915A base curve is allowed to degrade up to 80° deviation on a smooth curve from a straight line over the middle 400 MHz of the band, and up to 130° over the remaining 50 MHz at each end, then the requirements of Dc(3) a 1, 2 and 3 above shall be changed as follows:

1 Maximum phase error band center $\pm 24^{\circ}$.

2 Maximum phase error band edges $\pm 60^{\circ}$.

3 Average phase error band center $\pm 10^{\circ}$.

(4) Instrumentation shall be provided for remotely selectable readings of various operating DC voltages and currents of the driver and transmitter tubes. This facility shall allow monitoring of RF power metering couplers at interstage points in the driver chain in the presently existing operations room. Monitoring signals which go to an oscilloscope will be free from interference between different channels.

(5) Electronic means shall be provided for measurement and display in the control room of the phase versus frequency power amplifier output over the entire active bandwidth (250 MHz or 500 MHz).

(d) Remote Control of the wideband transmitter from the control console in the existing operations building is required. The functions requiring control are:

(1) Complete startup sequence procedure with indications of status, coolant flow, interlocks, magnets, filaments, cabinets, air flow where required and overload.

(2) Indication shall be provided for counting numbers of crowbar operations.

(3) Indication of power amplifier beam current, voltage, body current and average output power shall be continuously displayed at the control console.

(4) Capability for fine adjustment of magnetic fields of focusing magnets shall be included in the remote control position. Selectable current readings of the coils shall be provided.

(5) When the transmitter trips off, an indication shall be displayed at the remote control point of which item in the system initiated the action of circuit breaker opening.

(6) Facilities shall be provided at various critical locations throughout the entire system to initiate emergency shutdown action.

(7) Accessibility to the equipment components for test and repair must be considered in the design of the transmitter. All high voltages must be located behind interlocked doors, panels, or in tanks so as to allow limited tests or observations in the transmitter room located behind the antennas.

(e) Adaptation and Interface - The contractor shall employ existing facilities to the maximum extent. Drawing of enclosures, cable terminals and routings and hydraulic lines will be provided by the contracting agency as part of this Statement of Work.

E. PHASE II BROADBAND AMPLIFIER (VA915A)

The objective of this effort is to develop a broadband S-Band amplifier for operation in the Signal Processing Test Facility at RADC. The amplifier will replace the VA145 Twystron in the SPTF with a minimum of modification. The tube will be procured along with a suitable solenoid and socket modification kit. In the design of the tube, emphasis will be placed on smooth phase frequency and amplitude frequency response.

Three each Developmental Models of a tube having the following simultaneous characteristics;

(a) Center Frequency	3.35 Gigahertz (GHz)
(b) Minimum Instantaneous Bandwidth	(constant drive)
(1) 500 Megahertz (MHz)	-2 db (min) -6 db (max)
(2) 325 MHz	-0 db (min) -2 db (max)

(c) Peak Power Output. When a linear frequency modulation (FM) ramp is employed over the entire operating bandwidth during a single pulse, the mean peak power output over the pulse shall be a minimum of 7.15 megawatts. The maximum peakpower output shall not exceed 10 megawatts. The maximum peak power output shall occur within the range 3350 MHz $\pm$ 75 MHz.

(d) Duty; radio frequency (RF)	.0028
(e) Pulse length (RF)	40 microseconds (no detectable droop)
(f) Pulse length (video)	45 microseconds
(g) Gain	30 db (minimum)
(h) Pulsing	Cathode
(i) Mounting Position	Any

(j) Phase Linearity. The phase versus frequency characteristics shall show a deviation of not more than plus or minus 15 degrees from a smooth "base curve." The smooth "base curve" shall be continuous across the 500 MHz operating band with no reversal in the slope of the phase versus frequency characteristic describing the curve and with no more than two cycles of deviation from a linear curve of phase versus frequency. Further, the smooth base curve shall show a deviation of less than 50 degrees from a linear curve of phase versus frequency across the 500 MHz operating band and less than 30 degrees over the central 400 MHz of the operating band. The voltage sensitivity of the phase shall be less than 15 degrees per KV.

(k) Amplitude Response - The frequency versus amplitude response of the tube shall adhere as closely as possible to a smooth curve of one half cycle or less. Cyclic deviations from this smooth curve are particularly objectionable and shall be reduced to maximum extent possible. The sum of cyclic and random deviations from the smooth curve shall not exceed $\pm$ 0.5 db.

(l) Filament - The filament shall be so designed as to reduce phase and amplitude modulation to the maximum extent possible. DC filaments shall be employed.

(m) Focusing - Beam focusing shall be designed so that power supply ripple has minimum effect on phase and amplitude.

(n) Cooling Ethylene Glycol and Water (60-40).

(o) Shock - In accordance with Method 516.1 Procedure I basic design of MIL-STD-810B dated 23 June 1964.

(p) Vibration - In accordance with Method 514.1, paragraph 5, Figure 514-6 of MIL-STD-810B dated 23 June 1964.

(q) Acceleration: 5g continuous in any position.

(r) Spurious Radiation - The power level of any narrow band spurious oscillation integrated over a 3 MHz band shall be at least 60 db below the mean peak power output with linear FM ramp.

(s) Harmonics - Individual harmonics shall be at least 20 db below fundamental.

(t) Interchangeability - The tube shall be interchangeable with the VA145 Twystron with the minimum possible modification.

(u) Life - Each tube shall be warranted for a period of 500 hours of filament operation. Shelf life warrantee shall be 12 months.

(v) Environmental

(1) High and low temperature

a Nonoperating:	-60°C (-76°F) +60°C (+140°F)

(w) Position of operation - The tube shall be capable of meeting all specifications in any mounting position.

(x) RF Output Flange: Special Varian S-Band.

(y) Beam input conditions:

(1) Beam voltage (peak)	163.5 kv (min) 180.0 kv (max)
(2) Beam current (peak)	133 amps (min) 151 amps (max)
(3) Beam Perveance	1.8 micropervs (min) 2.0 micropervs (max)
(4) Beam impedance	1190 ohms (min) 1340 ohms (max)
(5) Maximum beam input (peak)	27.2 megawatts

(z) Two of the three tubes required shall meet all requirements of the specification. The third tube shall meet all requirements with the exception that at the contractors discretion the requirements for two of the four following items shall be reduced to 90% of the value stated.

(1) Power

(2) Phase

(3) Bandwidth

(4) Spurious Radiation

Two each interim tubes meeting all video and impedance requirements shall be supplied. The tubes shall be compatible with the peripheral equipment designed for the final tubes. These tubes need not meet the RF requirements set forth herein.

One set of hardware consisting of a solenoid and socket modification kit shall be supplied. The mechanical requirements of (o), (p) shall apply. The tube and socket shall be required to be completely operable on a non-stable platform varying continuously through angles up to 90 degrees in all directions. Two spare solenoids and other items subject to failure or breakage shall be supplied.

Test and Elevation shall be conducted as follows:

(a) Acceptance tests shall be conducted at the contractor's plant to demonstrate the requirements set forth in Items 1 and 2 by contractor personnel and witnessed by RADC's procuring activity. The test shall be conducted in accordance with the test procedures submitted under Data Item 6.3. All measurements shall be accurate within ± 2.5% or as accurately as the present state of art permits. The following test procedures shall apply to paragraph E(j).

(1) A dynamic slow frequency sweep method of measurement employing a precision pulse phase measurement system such as that provided in RANTEC Model ETS-17, similar WILTRON or equivalent equipment shall be used to obtain a continuous phase versus frequency plot across the operating frequency passband, and if the dynamic range of the equipment and the power output of the tube under test permits, plus 100 megahertz on either side of the band as well. The data shall be recorded by means of an X-Y Plotter. The pulse modulator, electromagnet power supplies, RF driver and other ancillary equipment shall provide a stable electrical environment with the following characteristics over at least 0.3 microseconds of the 45 microsecond video pulse:

a Modulator Pulse-to-Pulse Stability: Within ± 0.2%.

b Electromagnet Control Current: Within ± 0.1%.

c RF Drive Power: Constant within ± 0.3 db.

d Pulse Sample: Approximately 0.25 microseconds.

e Phase Measuring Equipment Error: Less than ± 2^{o}.

f Known Phase Measuring Auxiliary

Equipment Error: Not more than ± 5.0 degrees.

(2) The test shall be conducted as follows:

a A sufficient number of frequency sweeps shall be made for each individual tube to determine phase data containing unknown errors of electrical measurement. This data shall be averaged to obtain a curve of reference.

b The known errors of measurement shall be added to the reference curve to obtain a phase versus frequency envelope.

c The plus or minus 15 degrees permitted in the phase linearity specification (paragraph E(j)) shall be added to the envelope.

d The smooth "base curve" defined in the phase linearity specification shall fall within the phase versus frequency envelope defined by the above.

(3) Other tests to demonstrate compliance of the tubes to the specification shall be in conformance with MIL-E-1 where applicable.

F. DATA HANDLING EQUIPMENT

The main function of the data handling equipment will be to receive the time expanded output signal from the signal processor which represents the radar response from a 100 foot range window, and subsequently convert and digitally record the information.

The data handling equipment meeting the requirements may have a variety of forms, however, because of funding restrictions, emphasis will be placed upon those configurations which can accomplish the requirements at minimum cost. The maximum use shall be made of "off-the-shelf" items in order to insure minimum risk in attaining the desired goals. As a result of RADC's preliminary investigation it appears that sampling and A/D conversion may be the least expensive technique to provide the desired capability. The techniques that are presently envisioned are those which take advantage of the radar duty cycle (40 usec window at 70 pps), i.e., to buffer store and read out at a lower rate for recording the time expanded data with relatively inexpensive tape recorders.

This data handling equipment shall provide the necessary instrumentation for the tape recording of high precision radar signals and other related data which shall be used for further target analysis with a computer. The data handling equipment shall meet the following requirements:

(a) The DHE shall record the following analog input:

(1) Four sequential (in time) 40 usec blocks (Phase II information window) of analog signal representing the horizontal and vertical polarization amplitude and phase response of a 100 foot radar range window.

(2) Four sequential (in time) 20 usec blocks (Phase I) of information as in (1) above.

(b) The DHE shall not degrade the signal processor output and shall maintain the following parameters:

(1) Dynamic range 45 db.

(2) Video bandwidth (amplitude and phase channels) 2.5 MHz.

(3) Amplitude accuracy ± 1/4 db.

(4) Interblock relative amplitude accuracy ± 1/4 db.

(5) Intrablock relative amplitude accuracy ± 1/4 db.

(c) The DHE shall operate with a maximum of five volts peak signal input at 50 ohms.

(d) The information window rate shall be at 70 pps.

(e) The DHE shall record the following auxiliary data:

(1) Clock track and search.

(2) Radar range.

(3) Radar azimuth angle.

(4) Radar elevation angle.

(5) Frame synchronization.

(6) Frame identity.

(7) Doppler correction.

(8) Gain level.

(f) The output of the DHE tape recording system shall have the following characteristics:

(1) Digital recording and maximum use of IBM Format for direct computer usage.

(2) A minimum of 20 minutes recording time.

(3) Maximum tape utilization.

(4) Playback slow-down (a minimum of 4:1).

(5) Stabilities, skew, wow, flutter, etc., shall be such that recorded data can be related, e.g., the radar amplitude range response with its respective phase.

(6) Non-destructive readout.

(g) The DHE shall operate in the operations building at Floyd Test Site, New York, in conjunction with other operating equipment.

UNCLASSIFIED

Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

1. ORIGINATING ACTIVITY (Corporate author)
Rome Air Development Center
Surveillance and Control Division
Techniques Branch

402 452

2a. REPORT SECURITY CLASSIFICATION
Unclassified

2b. GROUP

3. REPORT TITLE

6 SIGNAL PROCESSING TEST FACILITY (PRELIMINARY DESIGN) Study).

[redacted]

9 Interim Report.

5. AUTHOR(S) (First name, middle initial, last name)
Techniques Branch, Surveillance and Control Division

[redacted]

11 Sep [redacted] 67

7a. [redacted]
12 133 p

7b. NO. OF REFS
12

[redacted]
16 AF 4506

[redacted]
17 450601

d.

9a. [redacted] REPORT NUMBER(S)
18 RADC 19 TR-67-256

9b. OTHER REPORT NO(S) (Any other numbers that may be assigned this report)

10. DISTRIBUTION STATEMENT
This document is subject to special export controls and each transmittal to foreign governments, foreign nationals or representatives thereto may be made only with prior approval of RADC (EMLI), GAFB, N.Y. 13440.

11. SUPPLEMENTARY NOTES
RADC Project Engineer
Thomas Maggio
AC 315 330-4026

12. SPONSORING MILITARY ACTIVITY
Rome Air Development Center (EMATS)
Griffiss Air Force Base, New York 13440

13. ABSTRACT

This report describes the preliminary design effort for a very wideband high power tracking radar. Consideration has been given to dispersion effects of radar components, phase and amplitude tolerances required to meet specified range side lobe levels, availability of components, etc. An approach for modifying an existing radar is also presented.

ejd

mlh

DD FORM 1 NOV 65 1473

UNCLASSIFIED

Security Classification

UNCLASSIFIED

Security Classification

14. KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
High Range Resolution						
Pulse Compression						
Tracking Radar						

UNCLASSIFIED

Security Classification